Billy Toman

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DEPARTMENT OF THE ARMY TECHNICAL MANUAL

F-M TRANSMITTERS AND RECEIVERS

DEPARTMENT OF THE ARMY • SEPTEMBER 1952

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CONTENTS

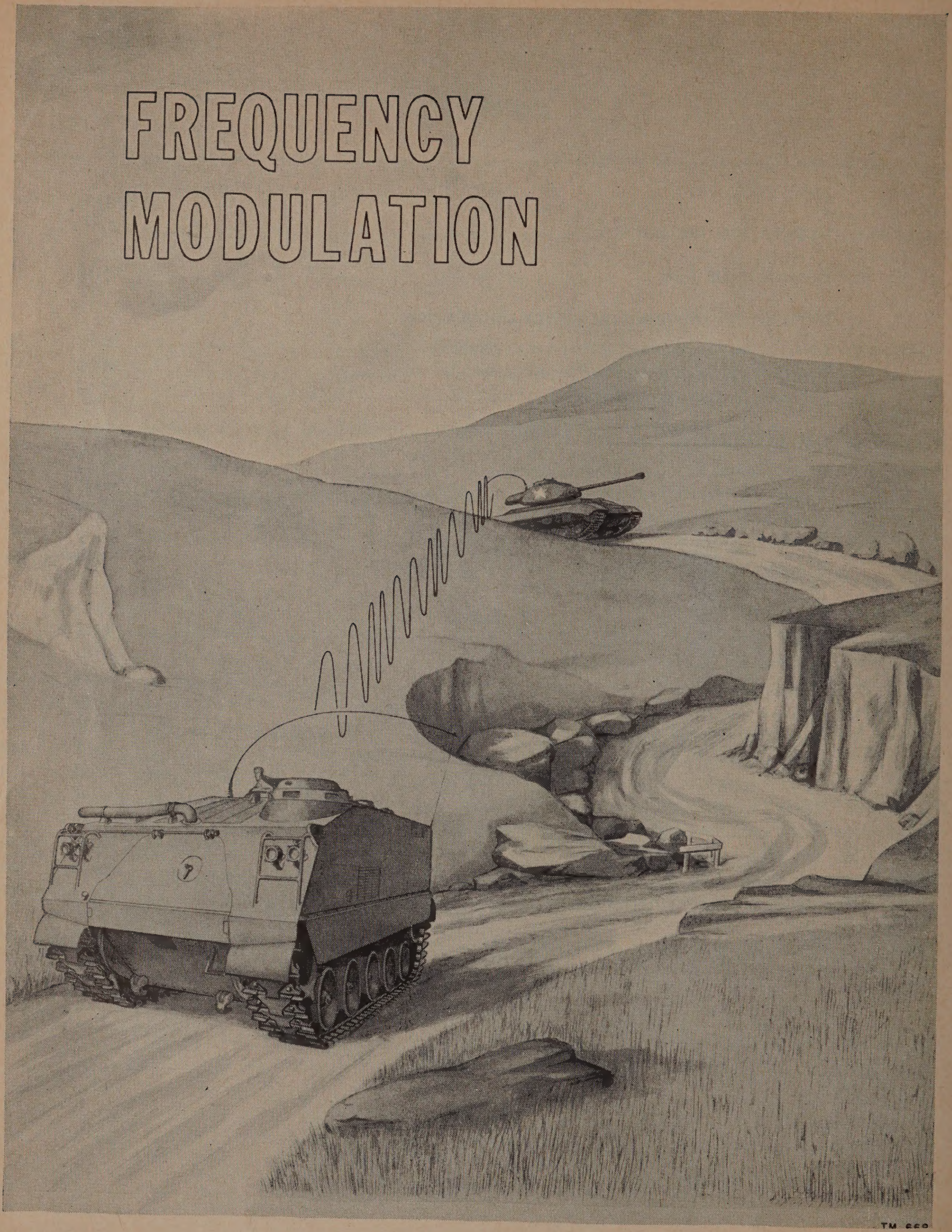


Figure 1. Mobile communication is an important function of f-m equipment.

CHAPTER 1

MODULATION SYSTEMS

1. Introduction

a. In the armed services, a reliable radio communication system is of vital importance. The swiftly moving operations of modern armies require a degree of coordination made possible only by radio. Today, the two-way radio is standard equipment in almost all military vehicles, and the walkie-talkie is a common sight in the infantry. Until recently, a-m (*amplitude modulation*) communication was used universally. This system, however, has one great disadvantage: Random noise and other interference can cripple communication beyond the control of the operator. In the a-m receiver, interference has the same effect on the r-f signal as the intelligence being transmitted because they are of the same nature and inseparable.

b. The engines, generators, and other electrical and mechanical systems of military vehicles generate noise that can disable the a-m receiver. To avoid this difficulty a different type of modulation, such as p-m (phase modulation) or f-m (frequency modulation), is used. When the amplitude of the r-f (radio-frequency) signal is held constant and the intelligence transmitted by varying some other characteristic of the r-f signal, some of the disruptive effects of noise can be eliminated.

c. In the last few years, f-m transmitters and receivers have become standard equipment in the Signal Corps, and their use in mobile equipments exceeds that of a-m transmitters and receivers. The widespread use of frequency modulation means that the technician must be prepared to repair a defective f-m unit, aline its tuned circuits, or correct an abnormal condition. To perform these duties, a thorough understanding of frequency modulation is necessary.

2. Carrier Characteristics

The r-f signal used to transmit intelligence from one point to another is called the *carrier*. It consists of an electromagnetic wave having amplitude, frequency, and phase. If the voltage variations of an r-f signal are graphed in respect to time, the result is a waveform such as that in figure 2. This curve of an unmodulated carrier is the same as those plotted for current or power variations, and it can be used to investigate the general properties of carriers. The unmodulated carrier is a sine wave that repeats itself in definite intervals of time. It swings first in the positive and then in the negative direction about the time axis and represents changes in the amplitude of the wave. This action is similar to that of alternating current in a wire, where these swings represent reversals in the direction of current flow. It must be remembered that the plus and minus signs used in the figure represent direction only. The starting point of the curve in figure 2 is chosen arbitrarily. It could have been taken at any other point just as well. Once a starting point is chosen, however, it represents the point from which time is measured. The starting point finds the curve at the top of its positive swing. The curve then swings through 0 to some maximum amplitude in the negative direction, returning through 0 to its original position. The changes in amplitude that take place in this interval of time then are repeated exactly so long as the carrier remains unmodulated. A full set of values occurring in any equal period of time, regardless of the starting point, constitutes one *cycle* of the carrier. This can be seen in the figure, where two cycles with different starting points are marked off. The number of these cycles that occur in 1 second is called the *frequency* of the wave.

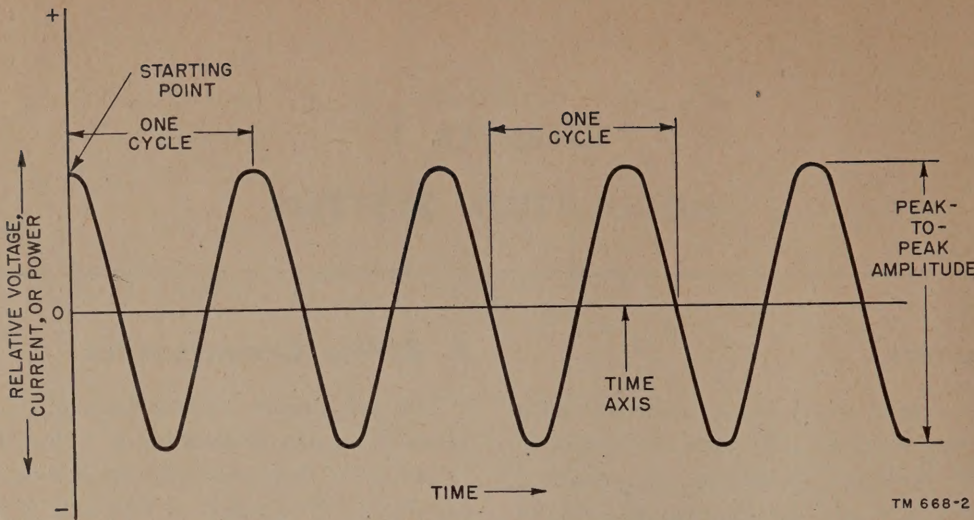


Figure 2. Graph of typical unmodulated carrier.

3. Amplitude Modulation

a. General. The amplitude, phase, or frequency of a carrier can be varied in accordance with the intelligence to be transmitted. The process of varying one of these characteristics is called *modulation.* The three types of modulation, then, are amplitude modulation, phase modulation, and frequency modulation. Other special types, such as pulse modulation, can be considered as subdivisions of these three types. With a sine-wave voltage used to amplitude-modulate the carrier, the instantaneous amplitude of the carrier changes constantly in a sinusoidal manner. The maximum amplitude that the wave reaches in either the positive or the negative direction is termed the *peak amplitude.* The positive and negative peaks are equal and the full swing of the cycle from the positive to the negative peak is called the *peak-to-peak amplitude.* Considering the peak-to-peak amplitude only, it can be said that the amplitude of this wave is constant. This is a general amplitude characteristic of the unmodulated carrier. In amplitude modulation, the peak-to-peak amplitude of the carrier is varied in accordance with the intelligence to be transmitted. For example, the voice picked up by a microphone is converted into an a-f (audio-frequency) electrical signal which controls the peak-to-peak amplitude of the carrier. A single sound at the microphone modulates the carrier, with the result shown in figure 3. The carrier peaks are no longer constant in amplitude because they follow the instantaneous changes in the amplitude of the a-f signal. When the a-f signal swings in the positive direction, the carrier peaks are increased accordingly. When the a-f signal swings in the negative direction, the carrier peaks are decreased. Therefore, the instantaneous amplitude of the a-f modulating signal determines the peak-to-peak amplitude of the modulated carrier.

b. Percentage of Modulation.

(1) In amplitude modulation, it is common

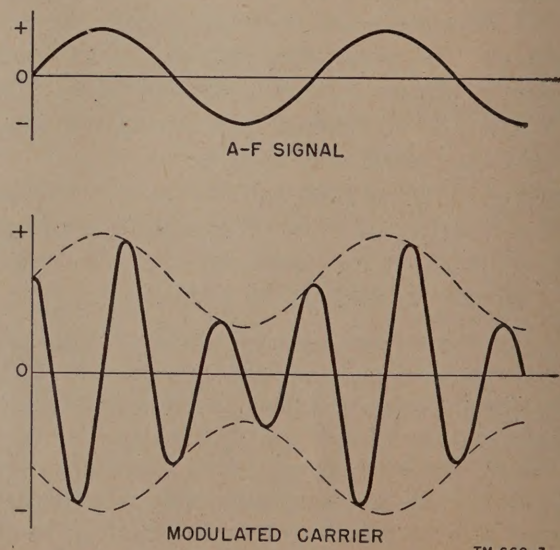


Figure 3. Effect of a-f signal on carrier in amplitude modulation.

practice to express the degree to which a carrier is modulated as a percentage of modulation. When the peak-to-peak amplitude of the modulating signal is equal to the peak-to-peak amplitude of the *unmodulated* carrier, the carrier is said to be 100 percent modulated. In figure 4, the peak-to-peak modulating voltage, E_A, is equal to that of the carrier voltage, E_R, and the peak-to-peak amplitude of the carrier varies from $2E_R$, or $2E_A$, to 0. In other words, the modulating signal swings far enough positive to double the peak-to-peak amplitude of the carrier, and far enough negative to reduce the peak-to-peak amplitude of the carrier to 0.

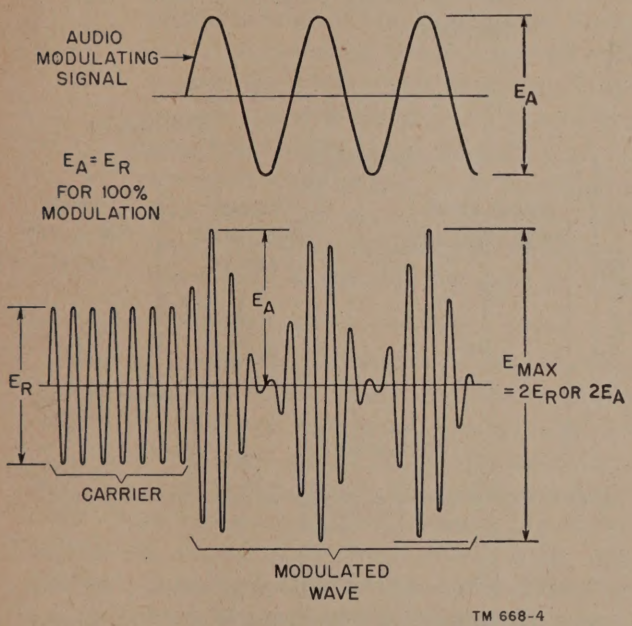


Figure 4. Carrier modulated 100 percent by a modulating signal.

(2) If E_A is less than E_R, percentages of modulation below 100 percent occur. If E_A is one-half E_R, the carrier is modulated only 50 percent (fig. 5). When the modulating signal swings to its maximum value in the positive direction, the carrier amplitude is increased by 50 percent. When the modulating signal reaches its maximum negative peak value, the carrier amplitude is decreased by 50 percent.

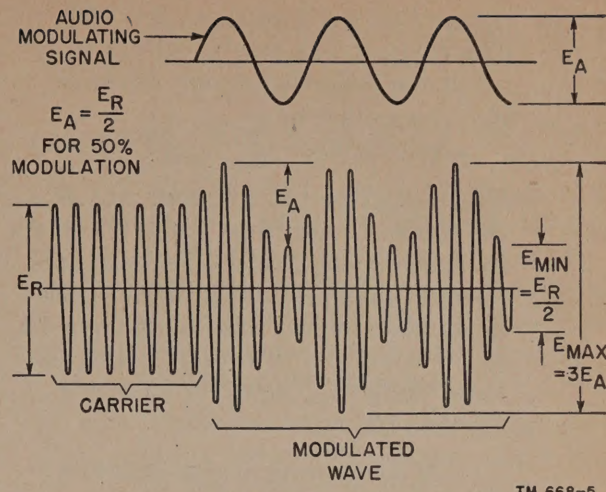


Figure 5. Fifty-percent modulation.

(3) It is possible to increase the percentage of modulation to a value greater than 100 percent by making E_A greater than E_R. In figure 6, the modulated carrier is varied from 0 to some peak-to-peak amplitude greater than $2E_R$. Since the peak-to-peak amplitude of the carrier cannot be less than 0, the carrier is cut off completely for all negative values of E_A greater than E_R. This results in a distorted signal, and the intelligence is received in a distorted form. Therefore, the percentage of modulation in a-m systems of communication is limited to values from 0 to 100 percent.

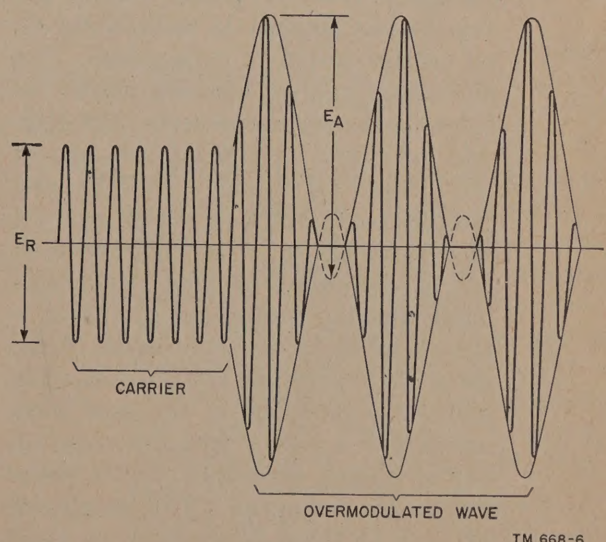


Figure 6. Overmodulation of carrier.

(4) The actual percentage of modulation of a carrier (M) can be calculated by using the following simple formula

$$M = \text{percentage of modulation} = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \times 100$$

where E_{max} is the greatest and E_{min} the smallest peak-to-peak amplitude of the modulated carrier. For example, assume that a modulated carrier varies in its peak-to-peak amplitude from 10 to 30 volts. Substituting in the formula, with E_{max} equal to 30 and E_{min} equal to 10,

$$M = \text{percentage of modulation} = \frac{30-10}{30+10} \times 100 = \frac{20}{40} \times 100 = 50 \text{ percent}$$

This formula is accurate only for percentages between 0 and 100 percent.

c. Side Bands.

(1) When the outputs of two oscillators beat together, or heterodyne, the two original frequencies plus their sum and difference are produced in the output. This heterodyning effect also takes place between the a-f signal and the r-f signal in the modulation process and the beat frequencies produced are known as *side bands*. Assume that an a-f signal whose frequency is 1,000 cps (cycles per second) is modulating an r-f carrier of 500 kc (kilocycles). The modulated carrier consists mainly of three frequency components: the original r-f signal at 500 kc, the sum of the a-f and r-f signals at 501 kc, and the difference between the a-f and r-f signals at 499 kc. The component at 501 kc is known as the *upper* side band, and the component at 499 kc is known as the *lower* side band. Since these side bands are always present in amplitude modulation, the a-m wave consists of a center frequency, an upper side-band frequency, and a lower side-band frequency. The amplitude of each of these is constant in value but the resultant wave varies in amplitude in accordance with the audio signal.

(2) The carrier plus the two side bands, with the amplitude of each component plotted against its frequency, is represented in figure 7 for the example given above. The modulating signal, f_A, beats against the carrier, f_C, to produce upper side band f_H and lower side band f_L. The modulated carrier occupies a section of the radio-frequency spectrum extending from f_L to f_H, or 2 kc. To receive this signal, a receiver must have r-f stages whose bandwidth is at least 2 kc. When the receiver is tuned to 500 kc, it also must be able to receive 499 kc and 501 kc with relatively little loss in response.

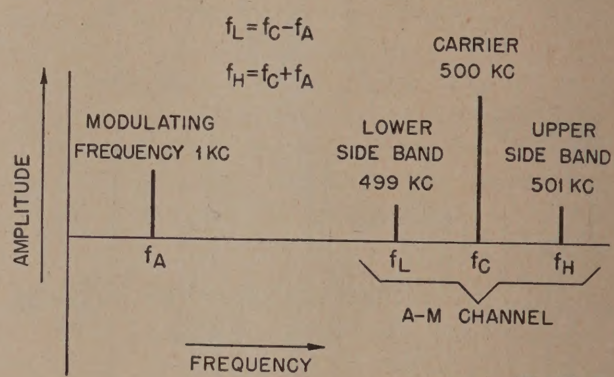


Figure 7. Frequency spectrum of amplitude-modulated wave.

(3) The audio-frequency range extends approximately from 16 to 16,000 cps. To accommodate the highest audio frequency, the a-m frequency channel should extend from 16 kc below to 16 kc above the carrier frequency, with the receiver having a corresponding bandwidth. Therefore, if the carrier frequency is 500 kc, the a-m channel should extend from 484 to 516 kc. This bandwidth represents an ideal condition; in practice, however, the entire a-m bandwidth for audio reproduction rarely exceeds 16 kc. For any specific set of audio-modulating frequencies, the a-m channel or band-

width is twice the highest audio frequency present.

(4) The r-f energy radiated from the transmitter antenna in the form of a modulated carrier is divided among the carrier and its two side bands. With a carrier component of 1,000 watts, an audio signal of 500 watts is necessary for 100-percent modulation. Therefore, the modulated carrier should not exceed a total power of 1,500 watts. The 500 watts of audio power is divided equally between the side bands, and no audio power is associated with the carrier.

(5) Since none of the audio power is associated with the carrier component, it contains none of the intelligence. From the standpoint of communication efficiency, the 1,000 watts of carrier-component power is wasted. Furthermore, one side band alone is sufficient to transmit intelligence. It is possible to eliminate the carrier and one side band, but the complexity of the equipment needed cancels the gain in efficiency.

d. Disadvantages of Amplitude Modulation. It was noted previously that random noise and electrical interference can amplitude-modulate the carrier to the extent that communication cannot be carried on. From the military standpoint, however, susceptibility to noise is not the only disadvantage of amplitude modulation. An a-m signal is also susceptible to enemy jamming and to interference from the signals of transmitters operating on the same or adjacent frequencies. Where interference from another station is present, the signal from the desired station must be many times stronger than the interfering signal. For various reasons, the choice of a different type of modulation seems desirable.

4. Phase Modulation

a. General.

(1) Besides its amplitude, the frequency or phase of the carrier can be varied to produce a signal bearing intelligence. The process of varying the frequency in accordance with the intelligence is frequency modulation, and the process of varying the phase is phase modulation. When frequency modulation is used, the phase of the carrier wave is indirectly affected. Similarly, when phase modulation is used, the carrier frequency is affected. Familiarity with both frequency and phase modulation is necessary for an understanding of either.

(2) In the discussion of carrier characteristics, carrier frequency was defined as the number of cycles occurring in each second. Two such cycles of a carrier are represented by curve A in figure 8. The starting point for measuring time is chosen arbitrarily, and at 0 time, curve A has some negative value. If another curve B, of the *same frequency* is drawn having 0 amplitude at 0 time, it can be used as a reference in describing curve A.

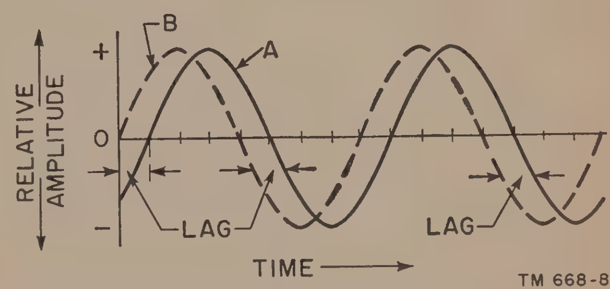


Figure 8. Determining relative phase from a curve of the same frequency.

(3) Curve B starts at 0 and swings in the positive direction. Curve A starts at some negative value and also swings in the positive direction, not reaching 0 until a fraction of a cycle after curve B has passed through 0. This fraction of a cycle is the amount by which A is said to *lag* B. Because the two curves have the same frequency, A will always lag B by the same amount. If the positions of the two curves are reversed, then A is said to *lead* B. The amount by which A leads or lags the reference is called its phase. Since the

reference given is arbitrary, the phase is relative.

b. Vector Representation.

(1) The cyclic changes of the carrier have been represented by plotting a curve of amplitude against time. It also is possible to represent the carrier cycle as the projection of a point rotating counterclockwise in a circle. This is called the vector representation and is accomplished by plotting the amplitude against the number of degrees of rotation of the point instead of directly against time. For each cycle of the carrier, the point rotates in one complete circle, or 360°. This is the *period* of the wave, or the time for one cycle.

(2) The carrier cycle as the projection of a point moving in a circle is plotted in figure 9. Starting from 0 the point rotates through 360° and back to 0, where the next cycle begins. When this point is projected along the set of axes on the right, one complete revolution of the point traces one complete cycle of the wave. The frequency of the wave in cycles per second is numerically equal to the revolutions per second made by the point. The relative phase of the wave shown is 0, since it has 0 amplitude at 0°. The peak amplitude of the wave is equal to the radius of the circle, and the peak-to-peak amplitude is equal to the diameter of the circle.

(3) The position of the point at any instant can be indicated by an arrow drawn from the center of the circle to the point. This arrow is called a *vector* and in the diagram the point is at a position of 45°. It must be remembered that this vector is not standing still, but is rotating at a frequency equal to that of the wave it represents. The vector is a convenient device with which to indicate phase relations between one wave and another or between a wave and an arbitrarily chosen reference. Suppose that a wave is observed staring at a time when its vector is in the 45° position (fig. 10). A second vector having 0 amplitude at 0 time and rotating in the same counterclockwise direction can be used as a reference for determining relative phase. In the graphic projection on the right, the two waves have the same frequency; however, the wave being observed is said to have a relative phase lead of 45°. An inspection of the figure shows that this is the central angle between the two vectors, as measured counterclockwise from the reference vector. The solid sine wave passes through 0 in a positive direction 45° ahead of the dotted reference sine wave shown at the right. If a particular carrier wave is considered in this manner, it has a relative phase relation at any instant to a reference carrier having the same frequency. The phase angle can be measured either in degrees or in radians. Since 360° equals 2π radians, the phase angle of 45° observed above can be expressed as $\pi/4$ radians, or simply $\pi/4$.

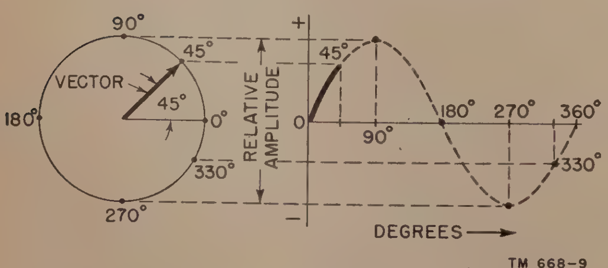


Figure 9. Projection of a point moving in a circle to form one cycle of a sine wave.

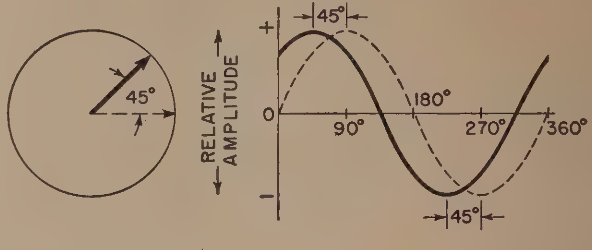


Figure 10. Measuring relative phase angle by means of vectors.

c. Phase Modulation.

(1) In phase modulation, the relative phase of the carrier is made to vary in ac-

cordance with the intelligence to be transmitted. The carrier phase angle, therefore, is no longer fixed. The amplitude and the average frequency of the carrier are held constant while the phase at any instant is being varied with the modulating signal (fig. 11). Instead of having the vector rotate at the carrier frequency, the axes of the graph can be rotated in the opposite direction at the same speed. In this way the vector (and the reference) can be examined while they are standing still. In A of figure 11 the vector for the unmodulated carrier is given, and the smaller curved arrows indicate the direction of rotation of the axes at the carrier frequency. The phase angle, ϕ, is constant in respect to the arbitrarily chosen reference. Effects of the *modulating signal* on the relative phase angle at four different points are illustrated in B, C, D, and E.

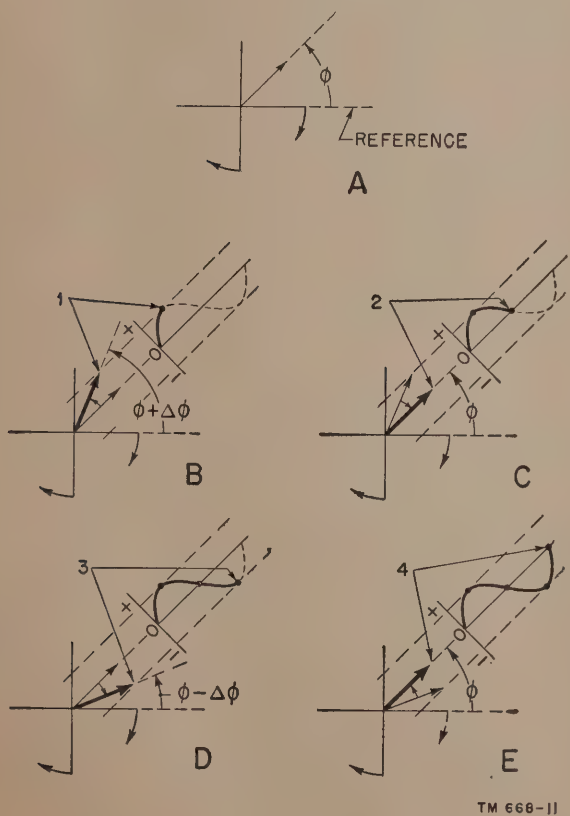


Figure 11. Successive vector representation of a phase-modulated carrier.

(2) The effect of a positive swing of the modulating signal is to speed the rotation of the vector, moving it counterclockwise and increasing the phase angle, ϕ. At point 1, the modulating signal reaches its maximum positive value, and the phase has been changed by the amount $\triangle\phi$. The instantaneous phase condition at 1 is, therefore, $(\phi + \triangle\phi)$. Having reached its maximum value in the positive direction, the modulating signal swings in the opposite direction. The vector speed is reduced and it appears to move in the reverse direction, moving toward its original position. When the modulating signal reaches 0 (position 2 in C), the vector has returned to its original position. The phase angle again is ϕ in respect to the reference. The modulating signal continues to swing in the negative direction and the vector is carried past its original position in a clockwise direction. When it reaches its maximum negative position, at 3 in D, the vector phase angle has been changed by $-\triangle\phi$, and the instantaneous phase angle is reduced to $(\phi - \triangle\phi)$. Finally, the vector returns to its original phase angle as the modulating signal amplitude falls to 0, at position 4 in E. The phase angle again is ϕ, and the cycle is complete. The entire cycle of phase shift is repeated for each cycle of modulating signal; that is, the frequency of the modulating signal is reproduced as the repetition rate of the cycle of phase shift.

(3) For each cycle of the modulating signal, the relative phase of the carrier is varied between the values of $(\phi + \triangle\phi)$ and $(\phi - \triangle\phi)$. These two values of instantaneous phase, which occur at the maximum positive and maximum negative values of modulation, are known as the *phase-deviation limits.* The upper limit is $+\triangle\phi$; the lower limit is $-\triangle\phi$. The relations between the phase-deviation limits and the carrier vector are given in figure 12, with the limits of $\pm\triangle\phi$ indicated.

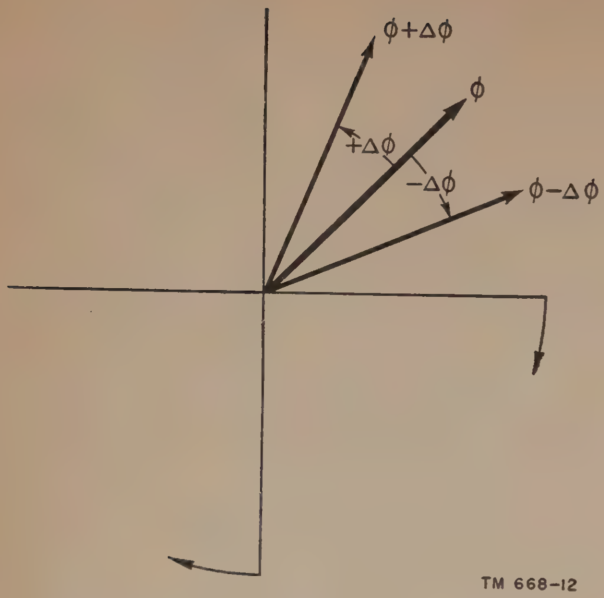


Figure 12. Phase-deviation limits of a modulated carrier.

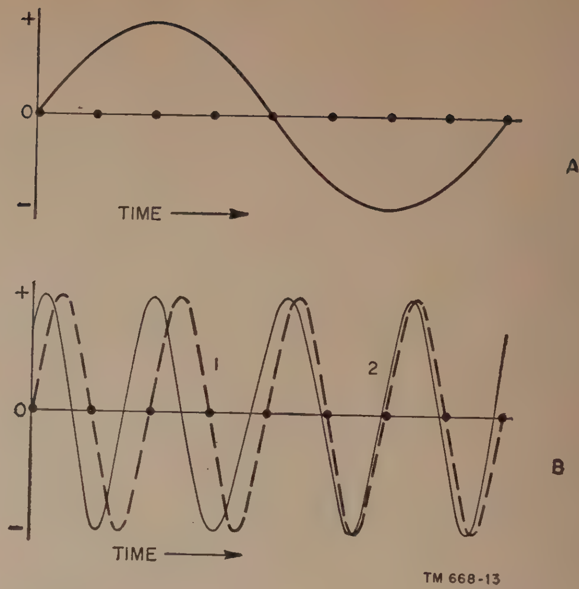


Figure 13. Phase-modulated carrier over 1 cycle of modulating signal.

(4) If the phase-modulated vector is plotted against time, the result is the wave illustrated in figure 13. The modulating signal is shown in A. The dashed-line waveform, in B, is the curve of the reference vector and the solid-line waveform is the carrier. As the modulating signal swings in the positive direction, the relative phase angle is increased from an original phase lead of 45° to some maximum, as shown at 1 in B. When the signal swings in the negative direction, the phase lead of the carrier over the reference vector is decreased to minimum value, as shown at 2; it then returns to the original 45° phase lead when the modulating signal swings back to 0. This is the basic resultant wave for sinusoidal phase modulation, with the amplitude of the modulating signal controlling the relative phase characteristic of the carrier.

d. P-M and Carrier Frequency.

(1) In the vector representation of the p-m carrier, the carrier vector is *speeded up* or *slowed down* as the relative phase angle is increased or decreased by the modulating signal. Since vector speed is the equivalent of carrier frequency, the carrier frequency must *change* during phase modulation. A form of frequency modulation, known as *equivalent f-m,* therefore, takes place. Both the p-m and the equivalent f-m depend on the modulating signal, and an instantaneous equivalent frequency is associated with each instantaneous phase condition.

(2) The phase at any instant is determined by the amplitude of the modulating signal. The instantaneous equivalent frequency is determined by the *rate of change* in the amplitude of the modulating signal. The rate of change in modulating-signal amplitude depends on two factors—the modulation amplitude and the modulation frequency. If the amplitude is increased, the phase deviation is increased. The carrier vector must move through a greater angle in the same period of time, increasing its speed, and thereby increasing the carrier frequency shift. If the modulation frequency is increased, the carrier must move within the phase-deviation limits at a faster rate, increasing its speed and thereby increas-

ing the carrier frequency shift. When the modulating-signal amplitude or frequency is decreased, the carrier frequency shift is decreased also. The faster the amplitude is changing, the greater the resultant shift in carrier frequency; the slower the change in amplitude, the smaller the frequency shift.

(3) The rate of change at any instant can be determined by the *slope*, or steepness, of the modulation waveform. As shown by curve A in figure 14, the greatest rates of change do not occur at points of maximum amplitude; in fact, when the amplitude is 0 the rate of change is maximum, and when the amplitude is maximum the rate of change is 0. When the waveform passes through 0 in the positive direction, the rate of change has its maximum positive value; when the waveform passes through 0 in the negative direction, the rate of change is a maximum negative value.

(4) Curve B is a graph of the rate of change of curve A. This waveform is leading A by 90°. This means that the

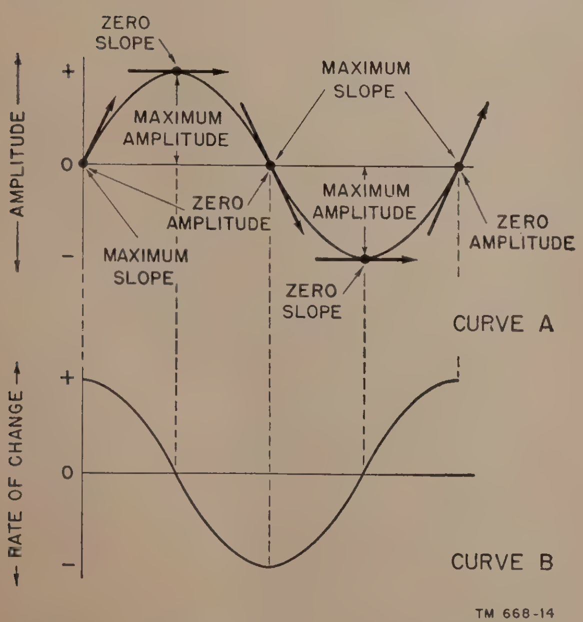


Figure 14. Amplitude compared to rate of change of a signal.

frequency deviation resulting from phase modulation is 90° out of phase with the phase deviation. The relation between phase deviation and frequency shift is shown by the vectors in figure 15. At times of maximum phase deviation, the frequency shift is 0; at times of 0 phase deviation, the frequency shift is maximum. The equivalent-frequency deviation limits of the phase-modulated carrier can be calculated by means of the formula,

$$\triangle F = \triangle\phi f \cos (2\pi \text{ ft})$$

where

$\triangle F$ is the frequency deviation,
$\triangle\phi$ is the maximum phase deviation,
f is the modulating-signal frequency,
$\cos$ $(2\pi$ ft) is the amplitude variation of the modulating signal at any time, t.

When $(2\pi$ ft) is 0 or 180°, the signal amplitude is 0 and the cosine has maximum values of +1 at 360° and —1 at 180°. If the phase deviation limit is 30°, or $\pi/6$ radians, and a 1,000-cps signal modulates the carrier, then

$$F = \frac{\pi}{6} \times 1{,}000 \times +1,$$
$$F = +523 \text{ cps, approximately.}$$

When the modulating signal is passing through 0 in the positive direction, the carrier frequency is raised by 523 cps. When the modulating signal is passing through 0 in the negative direction, the carrier frequency is lowered by 523 cps.

5. Frequency Modulation

a. When a carrier is frequency-modulated by a modulating signal, the carrier amplitude is held constant and the carrier frequency varies directly as the *amplitude* of the modulating signal. There are limits of frequency deviation similar to the phase-deviation limits in phase modulation. There is also an *equivalent phase* shift of the carrier, similar to the equivalent frequency shift in p-m.

b. A frequency-modulated wave resulting from 2 cycles of modulating signal imposed on

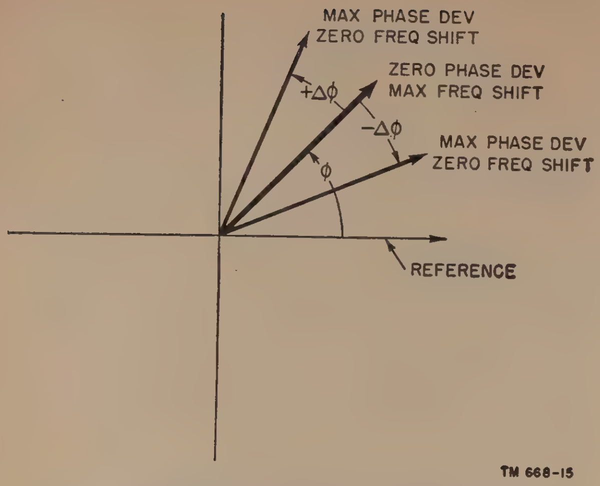


Figure 15. Phase deviation and frequency shift.

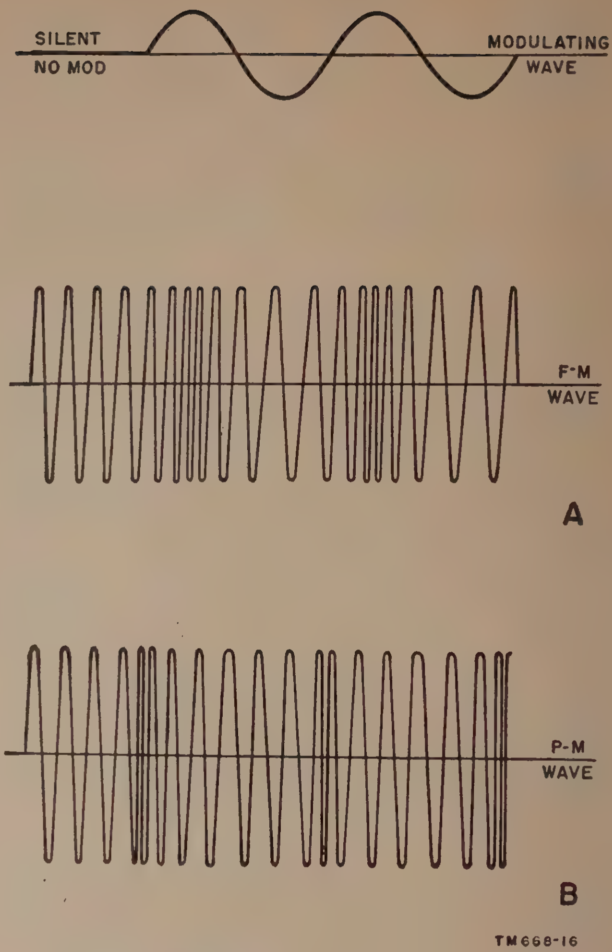


Figure 16. Comparison of f-m and p-m signals.

a carrier is shown in A of figure 16. When the modulating-signal amplitude is 0, the carrier frequency does not change. As the signal swings positive, the carrier frequency is increased, reaching its highest frequency at the positive peak of the modulating signal. When the signal swings in the negative direction, the carrier frequency is lowered, reaching a minimum when the signal passes through its peak negative value. The f-m wave can be compared with the p-m wave, in B, for the same 2 cycles of modulating signal. If the p-m wave is shifted 90°, the two waves look alike. Practically speaking, there is little difference, and an f-m receiver accepts both without distinguishing between them. Direct phase modulation has limited use, however, and most systems use some form of frequency modulation.

6. A-M, P-M, and F-M Transmitters

a. General. All f-m transmitters use either direct or indirect methods for producing f-m. The modulating signal in the direct method has a direct effect on the frequency of the carrier; in the indirect method, the modulating signal uses the frequency variations caused by phase-modulation. In either case, the output of the transmitter is a frequency-modulated wave, and the f-m receiver cannot distinguish between them.

b. A-M Transmitter.

(1) In the block diagram of the a-m transmitter (A of fig. 17), the r-f section consists of an oscillator feeding a buffer, which in turn feeds a system of frequency multipliers and/or intermediate power amplifiers. If frequency multiplication is unnecessary, the buffer feeds directly into the intermediate power amplifiers which, in turn, drive the final power amplifier. The input to the antenna is taken from the final power amplifier.

(2) The audio system consists of a microphone which feeds a speech amplifier. The output of this speech amplifier is fed to a modulator. For high-level modulation, the output of the modulator is connected to the final amplifier (solid arrow), where its amplitude modulates the r-f carrier. For low-

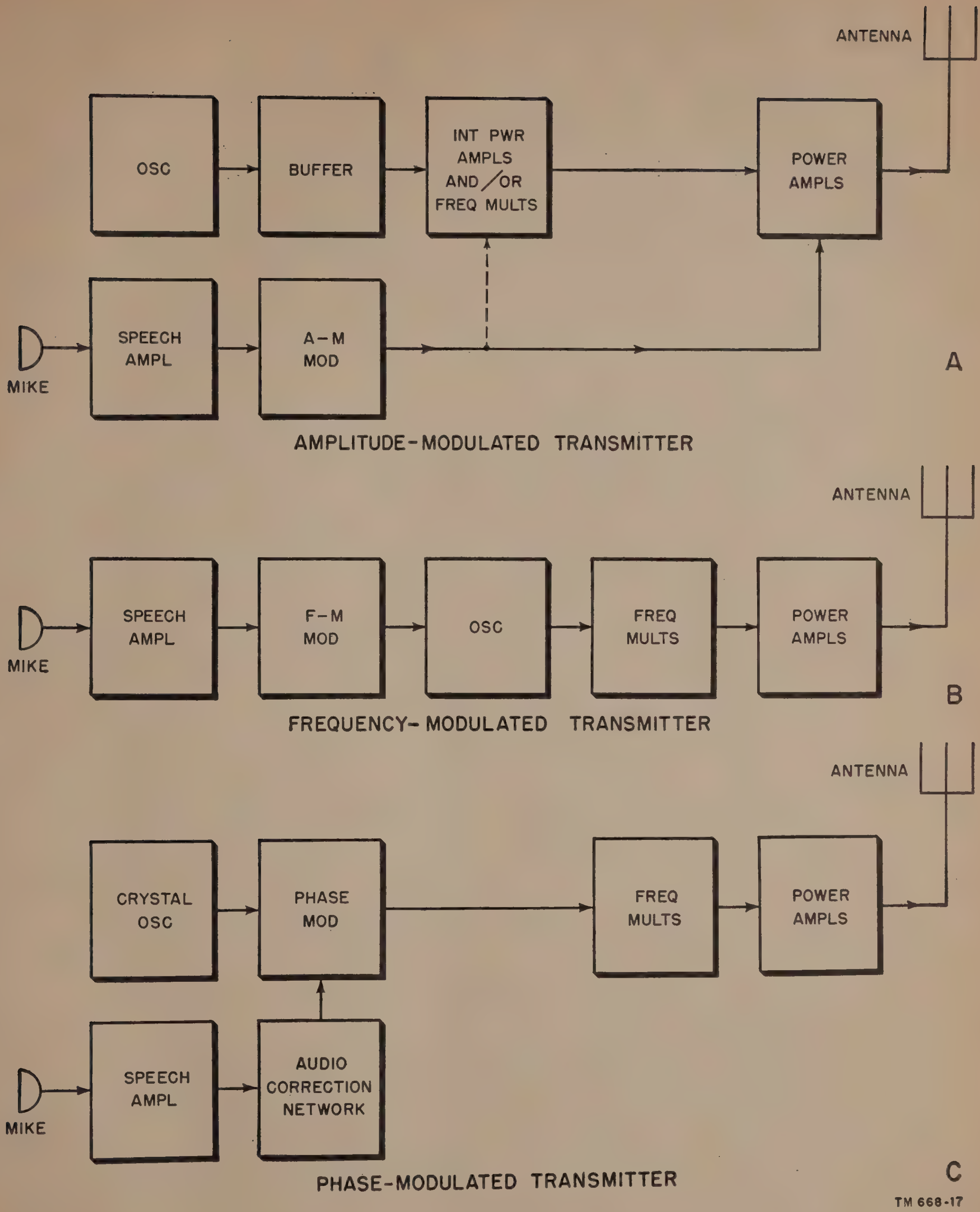


Figure 17. Basic a-m, p-m, and f-m transmitters.

level modulation, the output of the modulator is fed to the intermediate power amplifier (dashed arrow). The power required in a-m transmission for either high- or low-level modulation is much greater than that required for f-m or p-m.

c. P-M Transmitter. In the p-m, or indirect f-m, transmitter, the modulating signal is passed through some type of correction network before reaching the modulator, as in C. When comparing the p-m to the f-m wave, it was pointed out that a phase shift of 90° in the p-m wave made it impossible to distinguish it from the f-m wave (fig. 16). This phase shift is accomplished in the correction network. The output of the modulator which is also fed by a crystal oscillator is applied through frequency multipliers and a final power amplifier just as in the direct f-m transmitter. The final output is an f-m wave.

d. F-M Transmitter. In the f-m transmitter, the output of the speech amplifier usually is connected directly to the modulator stage, as in B. The modulator stage supplies an equivalent reactance to the oscillator stage that varies with the modulating signal. This causes the frequency of the oscillator to vary with the modulating signal. The frequency-modulated output of the oscillator then is fed to frequency multipliers which bring the frequency of the signal to the required value for transmission. A power amplifier builds up the signal before it is applied to the antenna.

e. Comparisons.

(1) The primary difference between the three transmitters lies in the method used to vary the carrier. In a-m transmission, the modulating signal controls the amplitude of the carrier. In f-m transmission, the modulating signal controls the frequency of the oscillator output. In p-m, or indirect f-m, transmission, the modulating signal controls the phase of a fixed-frequency oscillator. The r-f sections of these transmitters function in much the same manner, although they may differ appreciably in construction.

(2) The frequency multipliers used in a-m transmitters are used to increase the fundamental frequency of the oscillator. This enables the oscillator to operate at low frequencies, where it has increased stability. In f-m and p-m transmitters, the frequency multipliers not only increase the frequency of transmission, but also increase the frequency deviation caused by the modulating signal.

(3) In all three transmitters, the final power amplifier is used chiefly to increase the power of the modulated signal. In high-level a-m modulation, the final stage is modulated, but this is never done in either f-m or p-m.

7. A-M and F-M Receivers

a. General. The only difference between the a-m superheterodyne and the two basic types of f-m superheterodyne receivers (fig. 18) is in the detector circuit used to recover the modulation. In the a-m system, in A, the i-f signal is rectified and filtered, leaving only the original modulating signal. In the f-m system, the frequency variations of the signal must be transformed into amplitude variations before they can be used.

b. F-M Receiver. In the limiter-discriminator detector, in B, the f-m signal is amplitude-limited to remove any variations caused by noise or other disturbances. This signal is then passed through a discriminator which transforms the frequency variations to corresponding voltage amplitude variations. These voltage variations reproduce the original modulating signal. Two other types of f-m single-stage detectors in general use are the ratio detector and the oscillator detector, shown in C.

8. Summary

a. The carrier has the properties of frequency, amplitude, and relative phase.

b. In modulation, one of these three properties of the carrier is modified.

c. In amplitude modulation, the amplitude of the carrier is varied in proportion to the amplitude of the modulating wave.

d. The percentage of modulation of an a-m

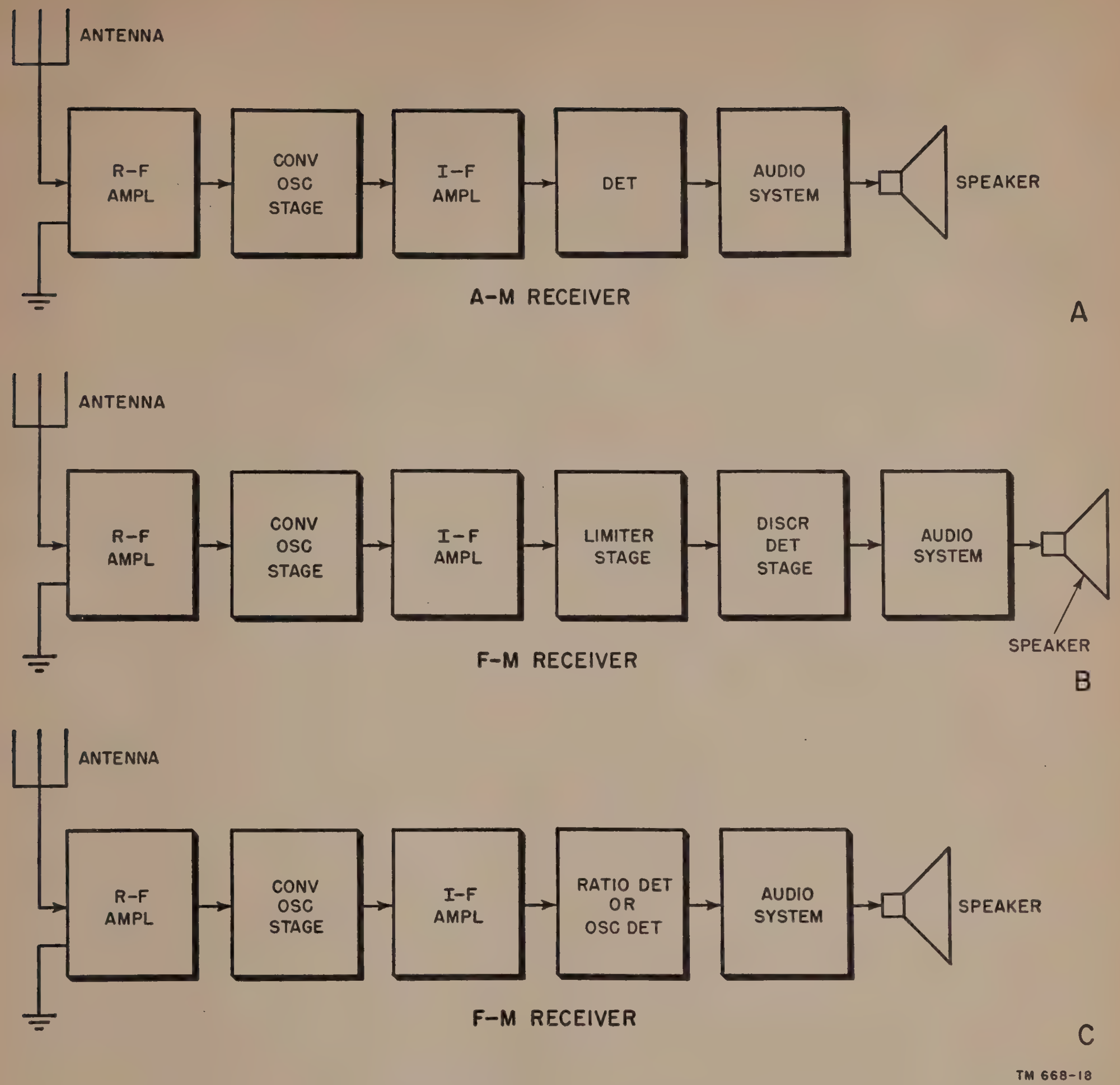


Figure 18. Basic a-m and f-m receivers.

wave is the percentage ratio of the peak-to-peak amplitude of the modulating voltage to the peak-to-peak amplitude of the carrier.

e. In amplitude modulation, the modulated wave consists of the carrier wave and of frequencies equal to the sum and difference between the carrier and the modulating frequency, called side bands.

f. The intelligence is contained in the side bands.

g. A-m has the disadvantage of being susceptible to some types of noise and interference.

h. In phase modulation, the instantaneous phase of the signal is varied by the modulating signal.

i. A change in phase is equivalent to an instantaneous change in frequency.

j. In a phase-modulation system, the equivalent frequency deviation is proportional to the frequency of the modulating signal.

k. When the carrier frequency is varied directly, the process is called direct f-m.

l. When the carrier frequency is varied indirectly, the process is called indirect f-m.

m. In a frequency-modulation system, the frequency varies directly with the amplitude of the modulating signal. The amplitude of the modulated wave remains constant, and the equivalent phase varies about a mean value.

9. Review Questions

a. What is meant by modulation of a carrier wave?

b. Name the most widely used types of modulation.

c. When a wave is amplitude-modulated 100 percent, what is the relationship between the amplitude of the modulating signal and that of the carrier?

d. What causes overmodulation of an a-m signal?

e. How many side bands are produced in a wave that is amplitude-modulated by a single sinusoidal tone?

f. What are the principal disadvantages of a-m?

g. Define phase deviation.

h. What happens to the instantaneous frequency during phase deviation?

i. How does the frequency of a p-m wave change during a cycle of modulating voltage?

j. How does the frequency of the f-m wave change during a single cycle of modulating voltage?

k. What characteristic of the modulating wave determines the maximum frequency deviation of an f-m wave?

CHAPTER 2

PRINCIPLES OF F-M

10. Frequency-Modulated Wave

a. Modulating Signal Amplitude and Frequency Deviation.

(1) In a frequency-modulated wave, the frequency varies instantaneously about the unmodulated carrier frequency in proportion to the amplitude of the modulating signal. When the modulating signal increases in amplitude, the instantaneous frequency increases, and when the modulating signal decreases, the frequency decreases.

(2) A radio-frequency carrier and an audio-frequency signal are shown separately in A and B of figure 19. When they are combined in the modulation process, the resultant signal is the f-m wave in C. As the amplitude of the audio signal increases in the positive direction, the modulated wave seems to bunch up, spreading out when the audio signal goes in the negative direction. These changes in the spacing of the modulated wave are caused by instantaneous changes in frequency. When the modulating signal is increased in amplitude, as in D, the changes in the spacing of the wave are proportionally greater, as in F. Therefore, the frequency deviation of the modulated wave is directly proportional to the amplitude of the modulating signal. When the audio voltage reaches its peak value in the positive direction, the frequency of the carrier is at its highest value above the center value. When the modulating voltage reaches its negative peak, the frequency of the carrier wave is reduced to its lowest value below that of the center carrier frequency. Maximum frequency deviation, therefore, takes place at the peaks of the audio signal.

b. Signal Frequency and Deviation. Figure 20 shows that each cycle of the modulating voltage, A, produces a corresponding variation in the frequency of the carrier wave, C. Two cycles of the audio wave produce 2 cycles of frequency change in the carrier. In D, the frequency of the modulating wave is increased so that in the *same* time interval the signal undergoes 3 complete cycles. Under this condition, the maximum frequency deviation remains the same as before; this is because the amplitude of the modulating wave has not been changed. The effective result of raising the frequency of the modulating signal is shown by comparing C and F. The audio signal is superimposed on the modulated carrier to demonstrate more clearly the relationship between the frequency of the modulating signal and the frequency changes in the carrier. This comparison indicates that, with a higher modulating frequency, the modulated wave deviates more frequently. However, the *limits* of frequency deviation are the *same,* regardless of the modulating frequency, since the audio signal is of constant amplitude.

c. Characteristics of F-M. The most important characteristics of a frequency-modulated wave are as follows: The amplitude of the modulated wave remains constant; the frequency of the modulated wave varies directly as the amplitude of the modulating signal; the limits of frequency shift on either side of the carrier are known as the *frequency deviation* limits. In an f-m system, the frequency of the modulating voltage determines the number of times per second that the frequency shifts between the

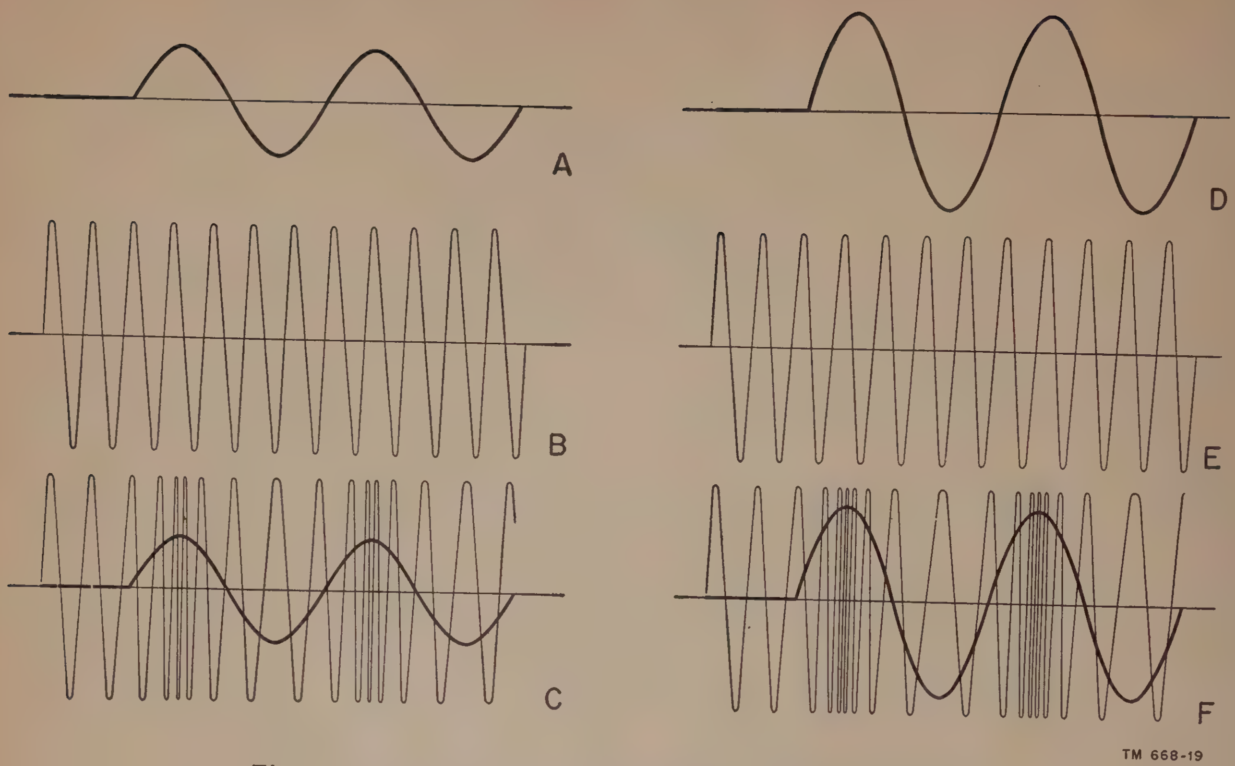


Figure 19. F-m waves for modulating signals differing in amplitude.

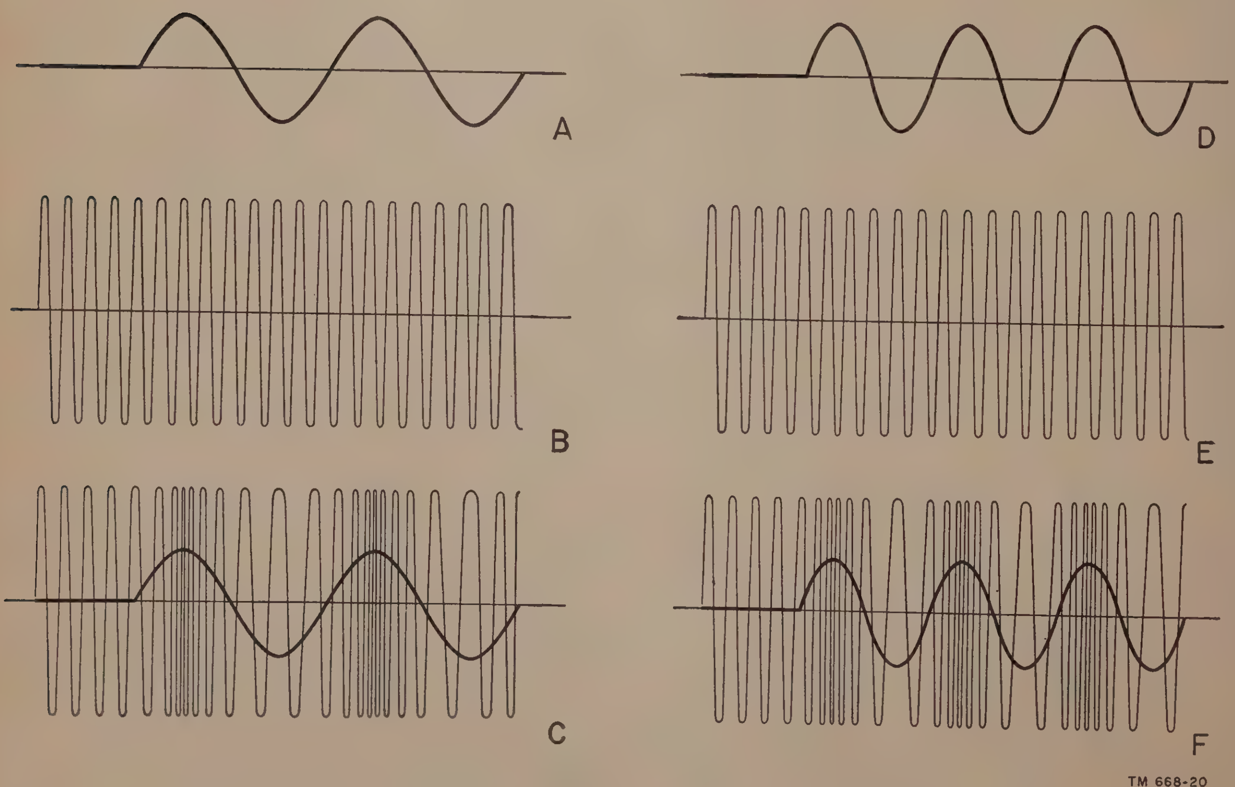


Figure 20. F-m waves for modulating signals differing in frequency.

deviation limits. The higher the frequency of the modulating signal, the greater number of times per second the frequency varies between the deviation limits set by the peak amplitude of the modulating signal. The ratio between the maximum frequency deviation and the maximum frequency of the modulating signal is called the *modulation index.*

$$\textit{Modulation index} = \frac{\textit{maximum frequency deviation}}{\textit{maximum frequency of modulating signal}}$$

d. Percentage of Frequency Modulation. The percentage of modulation of an f-m signal cannot be determined in the same manner as an a-m signal because 100-percent modulation would mean that the entire carrier varies in frequency from 0 to twice the carrier frequency. Percentage of modulation in f-m is defined as the percentage of maximum deviation incorporated in a transmitter for a particular type of service. For an f-m transmitter with maximum deviation of 75 kc, 100-percent modulation occurs when the transmitter deviates the full 75 kc. When the deviation falls to 37½ kc, the transmitter is being modulated only 50 percent. Such a definition is flexible, of course, and depends on the maximum deviation of the equipment used.

11. Side Bands

a. Amplitude Modulation.

(1) The intelligence superimposed on the carrier wave generates side-band frequencies closely adjacent to the carrier frequency. The generation of the side bands actually is the purpose of modulation, since the modulating energy cannot travel through space by itself. In an amplitude-modulated wave, the information represented by a sinusoidal modulating signal is carried in two side bands spaced on each side of the carrier frequency by an amount equal to the frequency of the modulating wave.

(2) If the modulating signal is composed of more than one sinusoidal wave, two side bands are generated for each sinusoidal component of the modulating wave. The greater the number of modulating frequencies present, the greater the number of side bands produced. However, only one pair of side bands is present for each frequency in the modulated wave, and the side bands farthest away from the carrier are those of the highest modulating frequency. Therefore, if the transmitter is amplitude-modulated no more than 100 percent, the bandwidth occupied by an a-m carrier and its side bands is twice the highest audio frequency in the modulating wave.

(3) In figure 21, frequency is plotted horizontally and the power contained in a signal of a particular frequency governs the vertical height. The carrier wave produced by the transmitter has only one frequency and its amplitude appears as a sharp line. Its power, represented by the heavy center line, is determined by the capabilities of the equipment. When a sine-wave signal from the audio power amplifier is combined with the carrier in the modulator stage, two identical side-band frequencies appear on either side of the carrier. The power of the applied audio signal is divided between them, and they are separated from the carrier by a frequency difference equal to the frequency of the audio signal. The carrier frequency of 1,000 kc is amplitude-modulated by a 5-kc tone, producing two side bands of equal power at 995 and 1,005 kc respectively, which are shown adjacent to the line representing the carrier. When a 10-kc modulating signal is added to the first, two additional side-band pairs appear, having frequencies of 990 and 1,010 kc, respectively.

b. Frequency Modulation.

(1) In an f-m wave, the amplitude of the modulating signal determines the departure of the instantaneous frequency from the center, or carrier frequency. The instantaneous frequency can be made to deviate as much as desired from the carrier frequency by changing the amplitude of the modu-

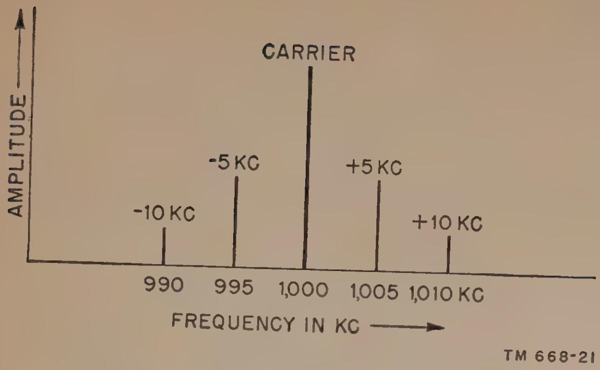


Figure 21. A-m carrier and side bands.

lating signal. It is possible to obtain a frequency deviation many times the frequency of the modulating signal itself. In practical equipment, this deviation frequency may be as high as several hundred kilocycles, even though the modulation frequency is but a few kilocycles. Therefore, the side bands generated by f-m are not restricted to the sum and difference between the highest modulating frequency and the carrier, as in a-m.

(2) Whereas in a-m, only two side bands, spaced equally on both sides of the carrier frequency, are generated, in f-m, many side bands are generated depending both in *number* and *amplitude* on the *modulation index*. The first pair of side bands in an f-m signal are those of the carrier frequency plus and minus the modulating frequency, and a pair of side bands will appear also at each multiple of the modulating frequency. As a result, an f-m signal occupies a greater bandwidth than does an a-m signal. For example, if a carrier of 1 mc (megacycle) is frequency-modulated by an audio signal of 10 kc, several side bands will be spaced equally on either side of the carrier frequency at 990 and 1,010, 980 and 1,020, 970 and 1,030, and so on. The total number present of *significant amplitude* (more than 1 percent of the amplitude of the unmodulated carrier) depends on the modulation index. With a high modulation index, more side bands are of appreciable amplitude, and the bandwidth is correspondingly greater.

c. Bandwidth. The maximum bandwidth of an amplitude-modulated transmission is twice that of the maximum frequency present in the modulating wave. Since the bandwidth in an f-m transmitter can exceed this by many times, the ratio of the bandwidth occupied to the absolute carrier frequency can be considerably larger than that of a-m transmitters. When a very wide bandwidth is used (wide-band f-m), it is necesary to choose a carrier frequency sufficiently high that the bandwidth is a small percentage of the carrier frequency, in order to permit a reasonable number of assigned channels. The f-m transmitter, however, can be adjusted so that the maximum bandwidth does not exceed that of an equivalent a-m transmission—that is, for the production of only one pair of side bands with significant amplitude. When the f-m transmitter is adjusted for a deviation that produces a bandwidth equal to that produced by an equivalently modulated a-m transmitter, it is called nfm (*narrow-band* f-m). With this narrow bandwidth, the transmitter can be operated at much lower frequencies (generally below 40 mc).

d. Relation Between Modulating Signal, Deviation, and Bandwidth. There are definite relations between the amplitude of the modulating signal, its frequency, the frequency deviation it produces, and the total bandwidth occupied by the resultant modulated f-m wave. If the frequency deviation is kept constant, *the number* of side bands *increases* as the modulating frequency *decreases*, and the total bandwidth occupied *decreases* as the modulating frequency *decreases*. The total bandwidth, however, can never be less than the bandwidth set by the peak-to-peak deviation alone, no matter how low the frequency of the modulating signal becomes. If the amplitude of the modulating signal *increases* and its frequency remains constant, the deviation *increases*, and the modulation index *increases*. This means that more energy goes into the outer side bands and correspondingly more of them increase to significant amplitude. The result is an increase in the number of useful side bands, as well as an increase in bandwidth.

e. Side-Band Spectrum and Modulation Index. The position of the side-band pairs for a single sinusoidal modulating wave depends only on the frequency of the modulating wave. The amplitude of the side bands depends on the ratio of the maximum frequency deviation of the carrier to the frequency of the modulating wave; that is, on the *modulation index.* The modulation index, in turn, depends on the amplitude of the modulating signal, because frequency deviation is proportional to signal amplitude. For a given modulation index and sinusoidal modulating frequency, the side-band pairs appear on either side of the carrier frequency (fig. 22). All the side-band components taken together form the *frequency spectrum* of the f-m wave. For example, if the modulating frequency is 15 kc and the frequency deviation is 75 kc, the modulation index will be 75/15, or 5, and the frequency components beyond the eighth pair of side bands will be less than 1 percent of the unmodulated carrier amplitude, and considered negligible.

f. Carrier Amplitude.

(1) The f-m wave consists of a center or carrier frequency and a number of side-band pairs, which, for a given audio frequency and amplitude, are constant. However, the resultant wave *varies in frequency* but is *constant in amplitude.* This resultant wave is the algebraic sum of the components which form it, and the carrier or center frequency will vary in amplitude with the modulation. When the transmitted signal is unmodulated, there is a certain constant amount of power in the carrier signal. When modulation is applied, power is taken from the carrier and forced into the side bands; therefore, the carrier amplitude, or center-frequency component, is reduced. The maximum power (fig. 22) is carried in the fourth side band, which is 4 times 15, or 60 kc, away from the carrier frequency.

(2) The carrier frequency or center-frequency component changes in amplitude with modulation, whereas, in a-m, the power for the side bands is supplied by the modulator and is not drawn away from the carrier. Since no information is in the carrier, reducing its amplitude increases the efficiency of operation in terms of power consumed. For some value of modula-

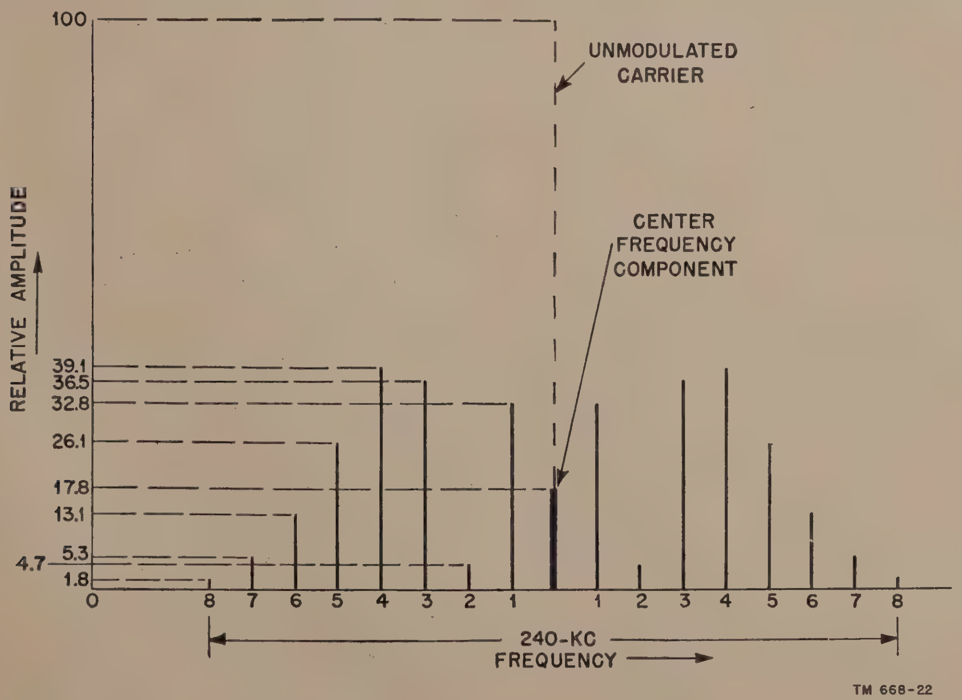


Figure 22. Frequency spectrum of f-m wave.

tion index and modulating frequency, carrier amplitude falls to 0 and all the power is contained in the side bands.

g. Numerical Values of Side Bands.

(1) In figure 22, with a frequency deviation of 75 kc and a modulating frequency of 15 kc, the center-frequency component is reduced to less than 20 percent of its unmodulated amplitude. If the modulating frequency is reduced to 5 kc with the same frequency deviation of 75 kc (fig. 23), the center frequency is reduced to only 1.4 percent of its unmodulated value. The side bands are spaced every 5 kc on either side of the center frequency out to the nineteenth pair of side bands, and all subsequent side bands are less than 1 percent of the unmodulated carrier amplitude.

(2) For the 15-kc modulating frequency of figure 22, with a modulation index of 5, the total bandwidth is 240 kc. In figure 23, with a 5-kc modulating frequency and a modulating index of 25, the total bandwidth occupied is 190 kc. The bandwidth of both is greater than the deviation limits of ±75 kc, which is equal to a peak-to-peak deviation of 150 kc. However, the side bands above or below the limit are relatively small in amplitude and can be disregarded. In both figures, the unmodulated carrier is shown in dashed lines for comparison with the amplitudes of the f-m side bands.

h. Side-Band Amplitude Computations.

(1) The relationship between the amplitudes of the side bands and an audio-modulating frequency for a modulation index of 2 is shown in figure 24. The deviation is 30 kc, with a modulating frequency of 15 kc. The modulated carrier wave, which is the resultant of the algebraic sum of the carrier and side bands is shown in A, with the modulating frequency, f_A, superimposed on it. At the positive peaks of the modulation cycle, the instantaneous frequency of the wave is f_C plus f_D, the frequency peak devia-

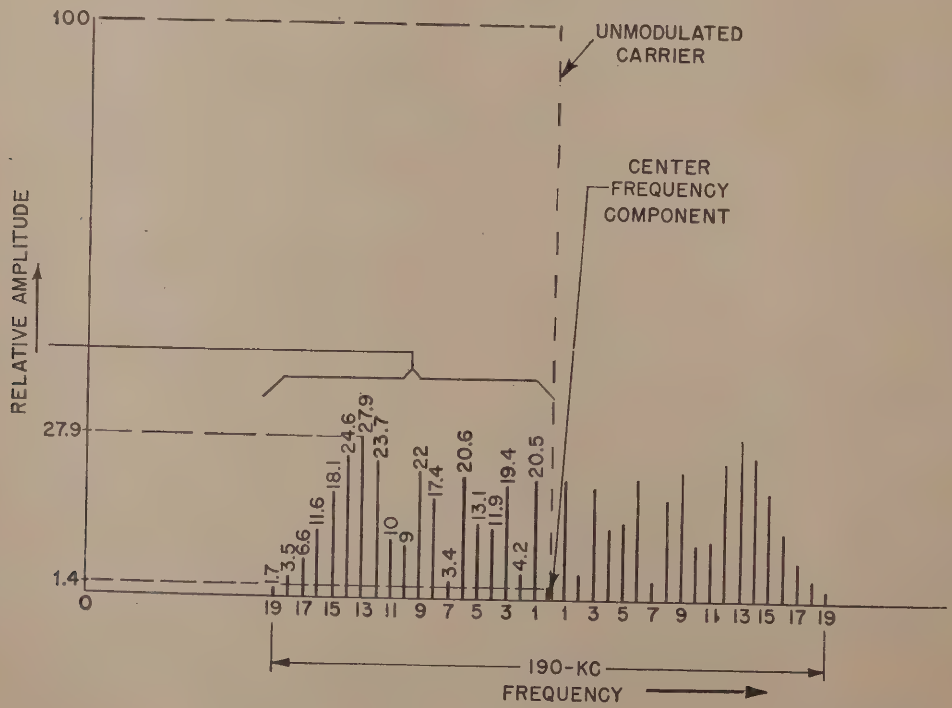


Figure 23. Another f-m frequency spectrum.

tion, whereas at the negative peaks the frequency is f_C minus f_D. The *peak-to-peak* deviation is, therefore, $2f_D$. If the carrier frequency is 100 mc with a frequency deviation of 30 kc, then the lower deviation limit is 99.97 mc and the upper one is 100.03 mc. There are four side-band pairs whose amplitudes exceed the 1-percent level and some of these are greater in amplitude than the center frequency. These are spaced on either side of the carrier frequency at intervals of 15 kc, as in C, D, E, and F, and each side-band pair has an amplitude as shown in G. The center frequency is reduced in amplitude to 22.4 percent of the unmodulated value. The first side-band pair at 99.985 mc and 100.015 mc has an amplitude of 57.7 percent of the unmodulated carrier value; the second side-band pair at 99.97 and 100.03 mc is 35.3 percent of the unmodulated carrier; and so on.

(2) To compute these bandwidths and side-band amplitudes, tables and graphs are available. Table I shows the number of *effective side-band pairs* for a given modulation index, and also the effective bandwidth that results for a given audio-modulating frequency, f_A. To compute the number of side-band pairs from the table, it is necessary to know the frequency deviation and the modulating frequency. The quotient of the two determines the modulation index, which in turn determines the number of effective side-bands pairs. For example, if the deviation is 25 kc and the modulating frequency is 5 kc, the modulation index is 5. From table I, a signal with a modulation index of 5 has eight effective side-band pairs, and the bandwidth is 16 times the modulating frequency of 5 kc, or 80 kc. The modulation indices that are less than .5 have only one pair of effective side bands and their effective bandwidths are all equal to $2f_A$. When using table I, if the modulation index is some fractional value, use the nearest

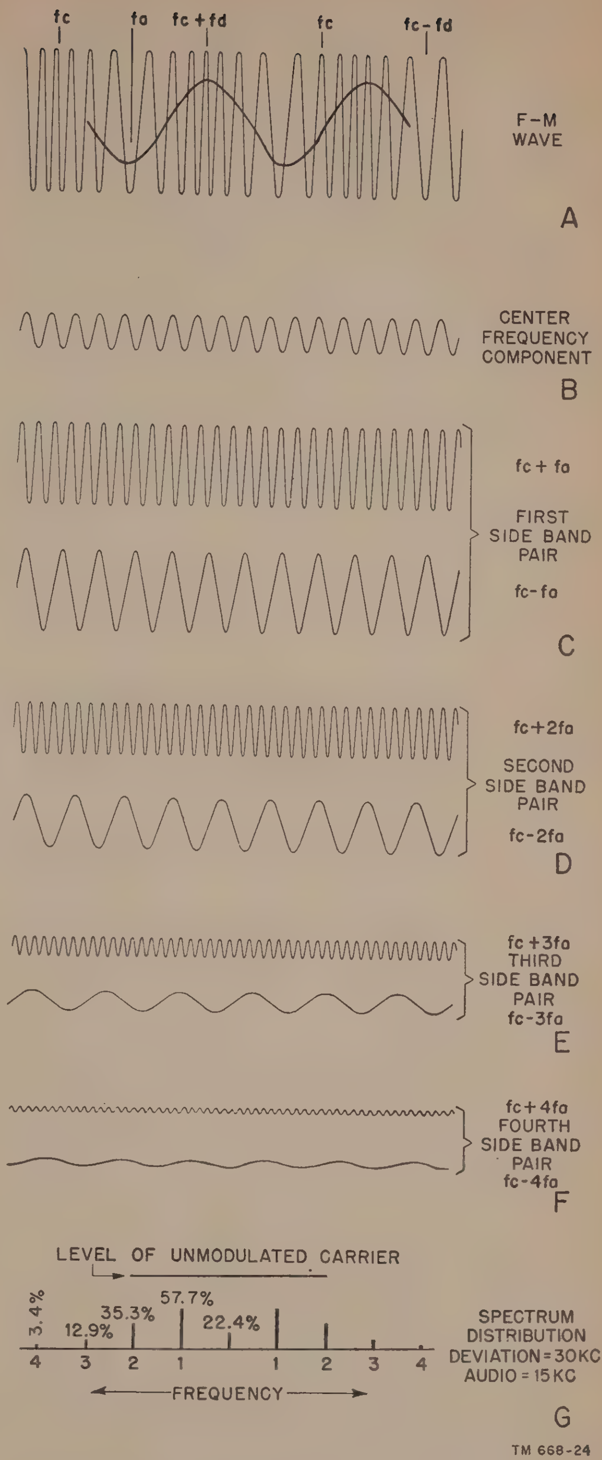


Figure 24. F-m carrier and side bands.

whole number. If the modulation index is 8¼, the number of effective side-band pairs for 8 is used; if it should be 8¾, the figure for 9 is used.

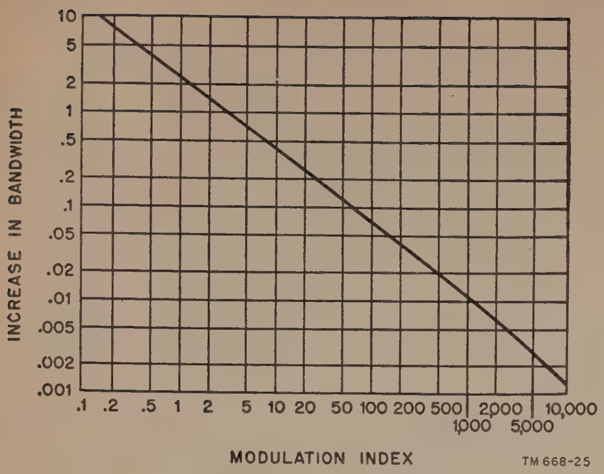


Figure 25. Bandwidth and modulation index.

Table I. Effective Bandwidth for Modulation Index

Modulation index $M = f_D/f_A$	Number of effective side-band pairs	Effective bandwidth
.5	2	$4f_A$
1	3	$6f_A$
2	4	$8f_A$
3	6	$12f_A$
4	7	$14f_A$
5	8	$16f_A$
6	9	$18f_A$
7	11	$22f_A$
8	12	$24f_A$
9	13	$26f_A$
10	14	$28f_A$
11	15	$30f_A$
12	16	$32f_A$
13	17	$34f_A$
14	18	$36f_A$
15	19	$38f_A$
16	20	$40f_A$
17	21	$42f_A$
18	23	$46f_A$
19	24	$48f_A$
20	25	$50f_A$
21	26	$52f_A$
22	27	$54f_A$
23	28	$56f_A$
24	29	$58f_A$
25	30	$60f_A$

(3) The same relationship is shown graphically in figure 25. The modulation index is plotted horizontally and the *increase* in bandwidth over the peak-to-peak deviation is shown vertically. A modulation index of 5 produces an increase of bandwidth of .6 or 60 percent. The peak-to-peak deviation is 2 times 25, or 50 kc. Sixty percent of 50 kc is 30 kc, and 30 kc plus 50 kc is equal to 80 kc. This is the same value as obtained with table I. The graph is especially useful in finding fractional values and modulation indices of less than one-half and more than 25.

i. Bandwidth and Guard Bands. The effective side bands must be at least as far from the carrier as the frequency deviation limits. Because of this, it is necessary to provide a channel or bandwidth that will handle the highest side-band component, plus a *guard band* that will absorb any side bands that extend beyond these limits. Since the modulating signal cannot always be specified and can vary over wide limits, it is easier to assign the channel in terms of deviation limits, and then to set aside some additional frequency space for guard bands on either side of the deviation limits. The channel with the guard bands assures a minimum amount of interference to stations operating on adjacent frequencies. In addition, the center carrier frequency of a particular station may vary from the assigned value and cause interference to stations on adjacent channels. The guard bands help to prevent this.

12. Types of Modulating Signals

a. Nonsinusoidal Modulating Waves. The preceding paragraphs are based on the assumption that the carrier wave was modulated by only one audio frequency that was sinusoidal in form. In general, waves containing information are neither sinusoidal nor composed of only one frequency. Moreover, the waves need not even be continuous; that is, they can start or stop abruptly. The modulation of the carrier by a nonsinusoidal wave can be understood by analyzing the character of the modulating wave. Waveforms can be analyzed as the sum of a number of sine waves of various amplitudes and frequencies. In respect to the frequency modulation of a carrier by a nonsinusoidal, discontinuous wave, the effective frequency variations follow the variations in the amplitude of the wave, just as they do in the sinusoidal wave. Therefore, the frequency deviation is still proportional to the peak amplitude of the signal,

although the variations are not as regular in a given time interval as they are in sine-wave modulation.

b. Speech Modulation. In normal speech, many different frequencies are present at the same time in the equivalent electrical wave. These frequencies range from 100 to 8,000 cps. For good intelligibility of speech only those between 300 and 3,000 cps need be transmitted. The average frequency content of speech depends on the voice of the speaker and on what he is saying; therefore, it is impossible to predict the side bands accurately. Standards must be set up that will duplicate the average speech spectrum, and a combination of 300 to 3,000 cps generally is used. The energy in high-pitched speech tones is much lower than in low-pitched speech, the amplitude of the 3,000-cps component being only one-tenth of the 300-cps components in a normal male voice. Therefore, the level of the 3,000-cps tone must be lower in amplitude to simulate the real qualities of speech. This means that the deviation for the high frequencies is correspondingly less, and the bandwidth is less than expected if it is assumed that equal intensities are used at 300 and 3,000 cps. Therefore, the bass frequencies cause the greatest deviation, even though the higher frequencies, if of equal intensity, tend to cause a wider bandwidth.

c. Pulse Modulation.

(1) When amplitude modulation with two audio frequencies is superimposed on the carrier, a pair of side bands is produced for every sine-wave component in the modulated wave. This fact can be demonstrated for as many components as are necessary to make up any given waveshape. A of figure 26 shows a rectangular wave composed of a large number of sine waves of various amplitudes and frequencies added together. Similarly, any other nonsinusoidal waveform can be analyzed in terms of a number of component sine waves.

(2) When nonsinusoidal waveforms amplitude-modulate a signal, they produce the same side bands that would be present if all of the sine waves to which they are equivalent were present. For example, rectangular amplitude modulation of a carrier wave produces side bands in symmetrical pairs about the carrier in relation to the amplitudes and frequencies of the sine-wave equivalent of the modulating waveform. This produces the spectrum of side bands shown in B. Each component making up the wave in A is plotted in terms of frequency and amplitude. This type of signal frequently is designated under a separate category as *pulse modulation.* Intelligence can be transmitted in such a system by controlling the amplitude,

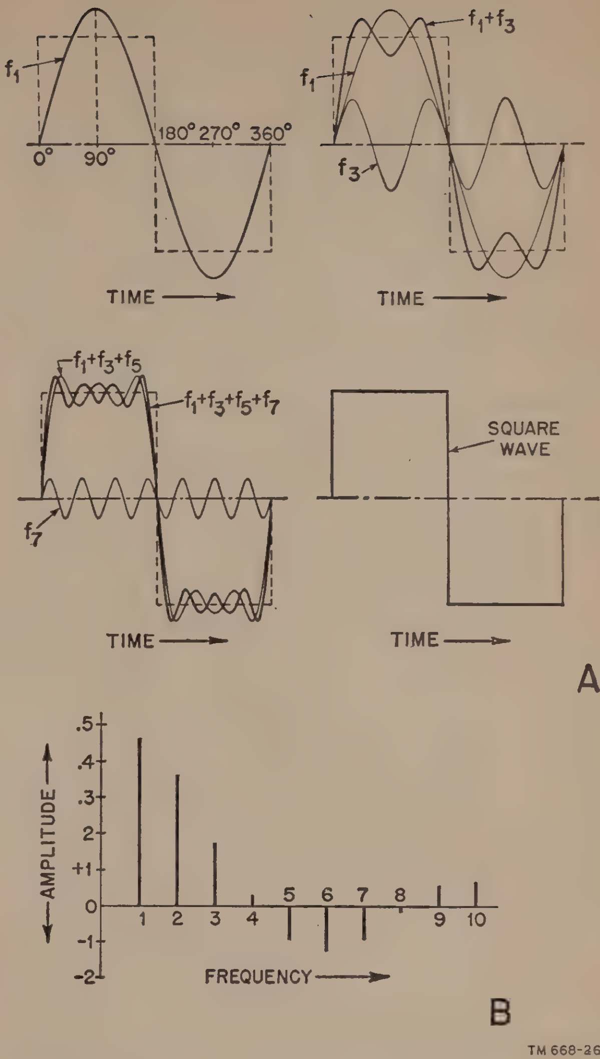


Figure 26. Sine-wave components of rectangular wave.

width, spacing, and position of the pulses. Frequency modulation, with the modulating wave in the form of rectangular pulses, is used in *frequency-shift* telegraphy. The frequency of the carrier is shifted abruptly between two values in accordance with the rectangular pulses received from a teletypewriter. Triangular pulses derived from certain types of facsimile transmitters for picture transmission are used also.

d. Side Bands—Complex Waves.

(1) When a sine-wave frequency modulates a carrier, many side bands are produced on either side of the carrier. These side bands are simple multiples of the modulating frequency with upper and lower side bands being identical in amplitude and *symmetrical* about the carrier.

(2) Assume that a signal of 5 kc modulates a carrier wave and produces 10-kc deviation, while at the same time a 10-kc signal modulates the carrier and produces a 10-kc deviation. The spectrum produced by the two signals tends to interact in a complicated way, depending on the amplitude and phase of the various side-band components. Not only the spectrum of each component is produced, but all possible combinations of each component with every other are present. Not all of these combinations, however, are of significant amplitude and the result is the production of the spectrum shown in figure 27. The actual instantaneous frequency deviation of the signal is the resultant of the sum of the two waves. That is, the two modulating waves add vectorially to produce a third that actually modulates the carrier, as shown in B.

(3) It can be seen that this spectrum is symmetrical about the carrier frequency but no longer corresponds to the spectrum of either component alone. Assume, however, that the deviation produced by the two component waves is not the same. In figure 28, a 5-kc signal produces a 30-kc deviation, and a 10-kc signal produces a 10-kc deviation. The modulating wave is shown with each of its components in B. The dotted line represents a 5-kc wave at three times the amplitude of the 10-kc wave. Since the deviation is proportional to the amplitude of the modulating signal, the amplitude of the wave producing 30-kc deviation is three times that of the wave producing 10-kc deviation. The result of combining the two waves is shown by the heavy black line, and it is this wave that causes the actual frequency deviation. The resultant spectrum is shown in A, and it is not symmetrical in respect to the carrier.

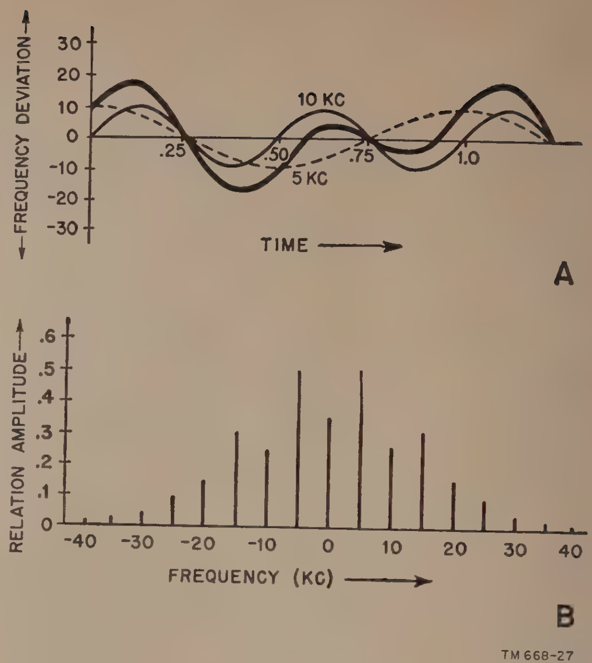


Figure 27. F-m for two modulating signals.

(4) In general, for a modulating signal that contains more than one sine wave, the side-band spectrum is not symmetrical about the carrier unless the modulating signal is symmetrical about the horizontal axis. For example, the wave (heavy line) in A of figure 27 is the same above and below the axis. This is not true of the wave in B of

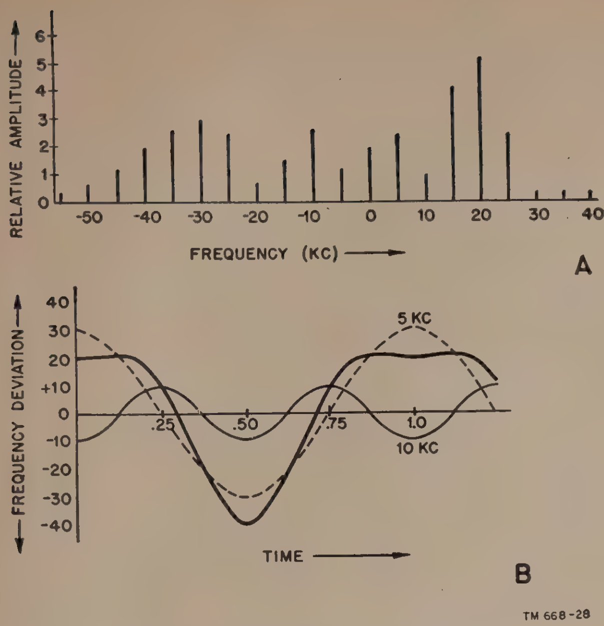


Figure 28. F-m for signals producing different deviations.

figure 28. The upper side band is associated with the shape of the positive half of the wave, whereas that of the lower side band depends on the negative half. This stems from the basic fact that an increase in amplitude of the modulating signal produced an increase in the instantaneous frequency of the carrier. Moreover, the energy in the side bands is distributed in accordance with the shape of the modulating wave. The effective bandwidth of the signal is increased, depending on whether the addition of the equivalent component sine waves of the complex waveform increases or decreases the peak deviation of the signal.

(5) Using the graph shown in C of figure 29, it is possible to determine the bandwidth of the signal with pulse modulation. For a wave with rectangular frequency modulation, shown in A, the spectrum is that in B. The maximum bandwidth required depends on the modulation index. This chart is used in the same manner as figure 25. When the deviation and the repetition rate of the rectangular pulses are known, the modulation index can be found by dividing the deviation by the repetition rate.

(6) For example, a frequency-shift telegraph transmitter deviation is 10 kc. (Each pulse of modulating signal shifts the frequency abruptly 10 kc). If the pulse repetition rate is 1,000 cycles per second, the modulation index is equal to

$$\frac{\textit{deviation}}{\textit{pulse repetition rate}} = \frac{10 \text{ kc}}{1 \text{ kc}} = 10$$

Entering the horizontal axis at a modulation index equal to 10, intersect with the curve at an increase in bandwidth of 2.25. The peak-to-peak deviation is 10 times 2 = 20 kc. The increase in bandwidth is, therefore, 20 times 2.25 = 45 kc, and the total band-

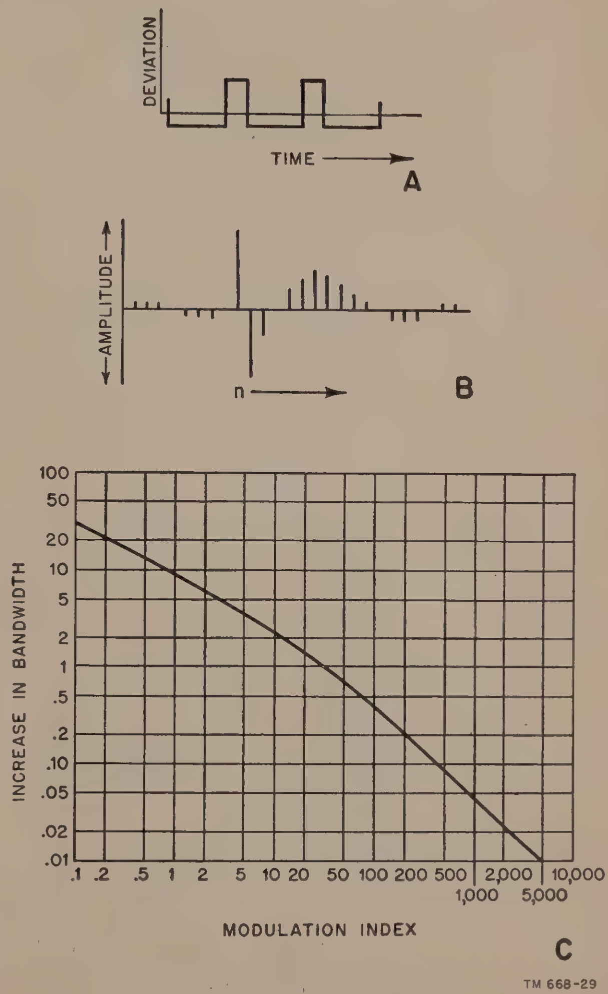


Figure 29. Rectangular modulation.

width is 20 plus 45 = 65 kc. The chart is computed with the pulse present one-quarter of the total time.

(7) If the modulating wave is triangular, as shown in A of figure 30, it produces the spectrum shown in B, and the total bandwidth can be calculated from the curves in C. This chart is used in the same manner as the one for the rectangular pulse. The modulation index is the quotient of the deviation and the repetition rate.

e. Bandwidth Limits for Nonsinusoidal Speech Modulation.

(1) The wide-band f-m wave contains numerous side bands that are distributed on either side of the carrier frequency. For sinusoidal modulating waves, the side bands are distributed equally above and below the carrier frequency. However, since most speech waves are not sinusoidal, the side bands in general are not of equal amplitude above and below the carrier. The number of side bands produced by a speech wave exceeds the number of sinusoidal components by a considerable amount. Moreover, the amplitude and the phase of each of these side bands vary in relation to the carrier frequency. Beyond certain frequency limits, however, the power contained in the outer side bands becomes very small, and so for all practical purposes they can be neglected. The point below which the side-band energy is negligible is arbitrary, depending on the kind of equipment that must be operated in the adjacent channel. As with sinusoidal modulation, when the the amplitude of the side-band component is less than 1 percent of the unmodulated carrier value, it is considered negligible.

(2) When the modulating signal is a mixture of many different waves distributed throughout the audio range, the ratio between frequency and phase deviation is hard to determine. The ratio of peak deviation in frequency to peak phase swing (*swing ratio*), measured in radians, gives some idea of the different conditions that prevail. This ratio, which is a constant for a simple sinusoid with a given modulation index, varies considerably with speech waves, depending on whether the modulation is indirect f-m or direct f-m. It is also dependent on the microphone, the characteristics of the audio-frequency amplifier, and how the total modulation applied to the transmitter is controlled to prevent overmodulation. In general, the swing ratio is greater than unity in an indirect f-m transmitter, whereas in a

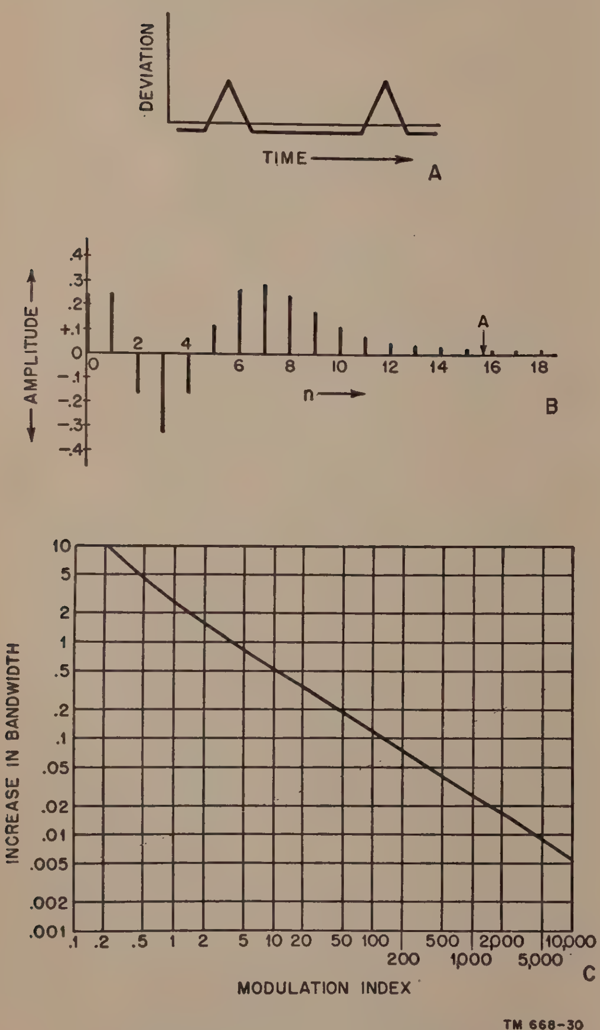


Figure 30. Triangular modulation.

direct f-m transmitter it is less than or equal to one.

(3) This means that in direct f-m transmitters the swing, or deviation in kilocycles, can be *less* for a given phase deviation than the expected value for a sine wave when speech is used. Conversely, the frequency deviation in an indirect f-m transmitter is *more* than the sine-wave value by as much as 50 percent, all depending on the character of the microphone, the audio amplifier, modulation limiting devices, and so forth. The importance of this relationship is that a directly modulated f-m transmitter has less peak deviation on speech waves than an indirect f-m transmitter when set for the same deviation, using a sine-wave source.

13. Preemphasis and Deemphasis

a. Preemphasis.

(1) In the transmitters used to convey speech, the deviation is the same for a given amplitude regardless of the frequency of the modulating signal. However, as signals pass through the transmitter, the receiver, and the space between them, certain amounts of unwanted noise and distortion are superimposed on the desired speech. This noise is distributed uniformly throughout the audible spectrum. Therefore, the ratio of the signal to the unwanted noise decreases in the higher frequencies because the speech amplitudes in this range do not have the intensity that the lower frequencies have. Moreover, the distortion increases in the high-frequency portion of the spectrum. The high frequencies make the greatest contribution to intelligibility of speech waves, since the consonants, which form the majority of speech sounds, have their peak energy in this part of the audio band.

(2) To avoid degrading the reproduction of consonants through poor signal-to-noise ratio in the upper end of the spectrum, a certain amount of added amplification (preemphasis) is provided for these frequencies. The result of this process should not sound unnatural when received, and the reverse procedure, *deemphasis,* therefore is used at the receiver. This combination of preemphasis and deemphasis provides a more uniform signal-to-noise ratio throughout the audio range. A transmitter using preemphasis has a wider side-band spectrum for speech than one without it. In general, the bandwidth of a speech signal deviating a transmitter 100 percent with preemphasis is about one-third greater than the deviation limits. If the deviation is 75 kc, for example, the total bandwidth at 100-percent modulation is about 200 kc (150 plus 50).

(3) The fact that preemphasis results in a greater bandwidth for a given deviation always must be taken into account. However, the possibility of overmodulation is not likely, since the high-frequency components of the signal originally are weak and the preemphasis merely brings them up to the level of the low tones. It does not cause overmodulation of an f-m transmitter, although the deviation limits set for the particular unit will be increased. It has been shown that the effective bandwidth increases as the audio frequency increases, and also that, as the upper audio frequency increases in level, more and more of the outer side bands rise above the 1-percent margin.

(4) The preemphasis characteristic of an f-m transmitter can be specified by a graph (fig. 31) showing the relationship between the audio input and the modulated output. The frequency of the audio spectrum is plotted horizontally, and the output of the unit for an input that is constant in respect to frequency is shown vertically. This curve shows that the output remains relatively constant from 50 to about 500 cps and then rises abruptly to a peak at 15,000 cps. Since this rise is specified in db (decibels), a change of

6 db means a doubling of the amplitude of the signal. Therefore, when the graph shows a rise of 18 db from 1,000 to 15,000 cps, it means that the amplitude has doubled three times. The resultant output at 15 kc, therefore, is 2 times 2 times 2, or 8 times the output at 1,000 cps.

b. Deemphasis. At the receiver, the reverse characteristic of preemphasis is used so that the natural balance between high and low frequencies in speech is not upset. The characteristics of preemphasis and deemphasis normally are achieved by simple electrical combinations of resistance, capacitance, and inductance connected to give the desired relationship between the input and the output voltages of the network. The characteristics of speech are complicated, and, therefore, the networks chosen represent a compromise between duplicating the exact loss of high frequencies and using as few parts as possible. In general, preemphasis and deemphasis circuits are very simple combinations of a capacitor and a resistor, or an inductor and a resistor.

c. Preemphasis Network. A simple preemphasis network consisting of an inductor and a resistor connected in the grid circuit of a vacuum-tube amplifier is shown in A of figure 32. In this circuit the audio voltage is impressed across the inductor and the resistor in series, and the output is taken across the inductor. Since the impedance of the coil rises with frequency, and the resistance remains constant, the voltage across the coil rises. The ratio of inductance to resistance determines the time constant of the combination, and the preemphasis characteristic can be specified completely in terms of the time constant. When the inductance is given in henrys and the resistance in megohms, the time constant is in microseconds. For example, calculate the time constant of the network (fig. 32) which contains a resis-

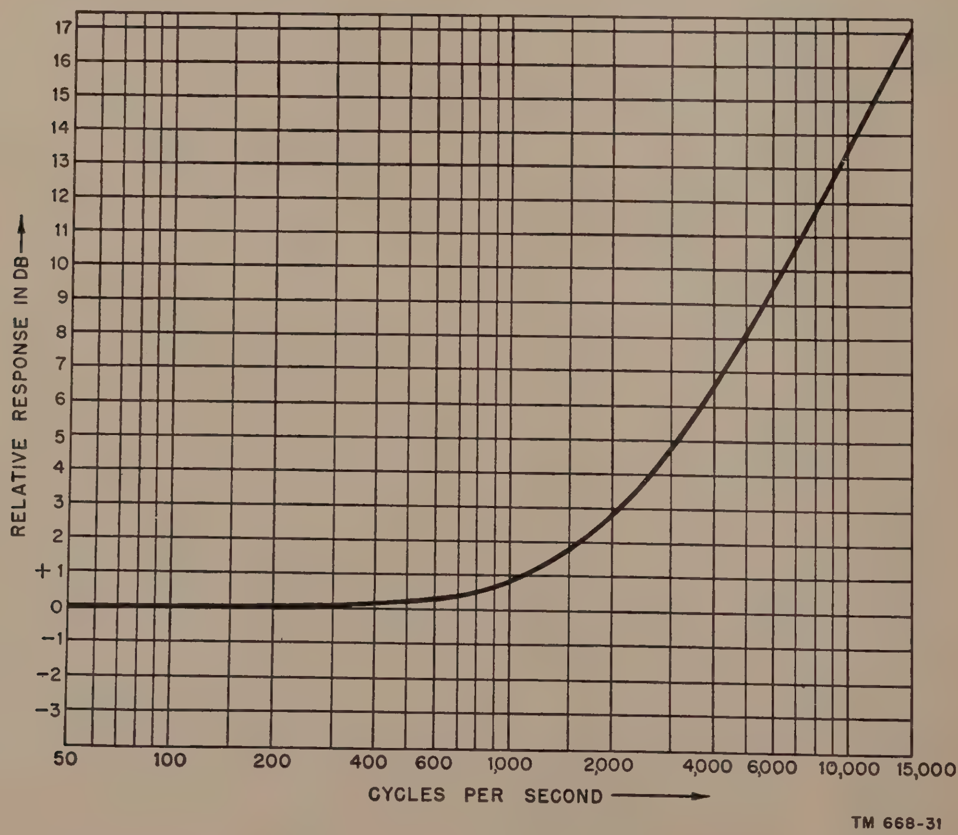
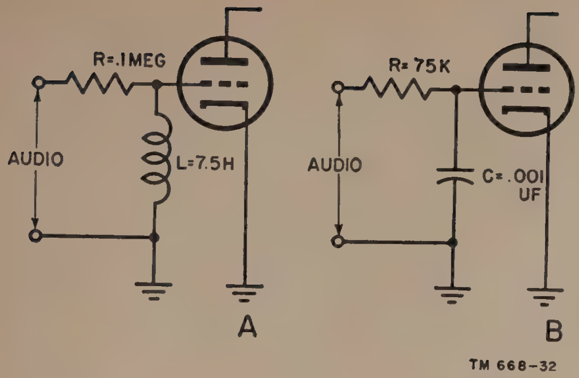


Figure 31. Preemphasis curve.



A. Preemphasis B. Deemphasis

Figure 32. Networks.

tance of 100,000 ohms (.1 meg) and an inductance of 7.5 henrys.

$$\text{Time constant} = \frac{L}{R} = \frac{7.5}{.1} = 75 \text{ microseconds}$$

For the specific ratio of inductance to resistance in figure 32, the graph of output voltage in respect to input voltage is shown in figure 31.

d. Deemphasis Networks. The deemphasis at the receiver must be the reverse of the preemphasis characteristic. This is accomplished by making the time constant of the resistor and capacitor in B of figure 32, equal to that of the preemphasis circuit. Since capacitive reactance decreases with increased frequency, the voltage across it decreases as the frequency rises. When the proper time constant is chosen, the higher frequencies are restored to their normal values. If the capacitor is in microfarads and the resistance is given in ohms, the product of R times C gives the time constant in microseconds. For example, in the circuit in B, the capacitor is .001 μf (microfarad) and the resistance is 750,000 ohms. What is the time constant?

$$\text{Time constant} = R \times C = 750{,}000 \times .001 = 75 \text{ microseconds}$$

This is the same time constant as that of the inductor and resistor in A.

14. Noise and Communication

a. General. One of the greatest disadvantages in transmitting information by amplitude modulation is the susceptibility of an a-m signal to both natural and man-made noises. Some of the disadvantages of a-m with regard to noise and freedom from interference can be overcome by using frequency modulation. Although f-m is not the most efficient system for overcoming noise, it is one of the easiest methods to apply. Modulation by means of short pulses of energy is more efficient, but the transmitters are much more complicated than those used for f-m. They seldom are employed except where large, cumbersome equipment can be used.

b. Impulse and Fluctuation Noise. Most man-made noises fall into two general classifications; *impulse noise* and *fluctuation noise*. *Impulse noise* consists of sharp pulses of r-f voltage which, when detected in a receiver, take the form of equally sharp pulses of audio voltage. They are often many hundreds of times greater in amplitude than the desired signal and make it impossible for the desired signal to be received. Perhaps the most common producers of impulse noise are the ignition systems of gasoline engines. Since many radio applications call for the installation of communication equipment in military vehicles, impulse noise is a problem. Steps are taken to eliminate as much of this noise as possible in military vehicles, but such elimination measures can never be perfect. There is always a residual component of the noise which may cause serious difficulties if the received signals are weak. The second kind of man-made noise, called *fluctuation noise,* is of a more continuous character. It appears as a broad band of many pulses which bear little or no relation to each other. Such noises are produced to a great extent by rotating electrical machinery, gas rectifiers, high-voltage transmission lines, and similar power devices. The noise from a small motor, although frequently weaker than the signal by a considerable amount, is capable of causing severe interference and possible interruption to a-m reception.

c. Noise Transmission. Man-made noise can reach the receiver in several possible ways. It can be received as a radiated signal along with the desired signal, or it can be picked up at the input of the receiver by transfer of energy through the capacitance between the antenna and the noise-producing device. Power lines in the vicinity of the antenna to which a noise-producing device is connected may induce a

noise directly in the antenna, or the noise can be transmitted directly over the power line to the receiver itself.

d. Man-Made Noise Frequencies. Man-made noise is not distributed uniformly through the spectrum. Impulse noise is most bothersome in the frequencies approximately from 15 mc to 160 mc and fluctuation noise generally is more severe at lower frequencies, and with a maximum intensity reached at frequencies well below 20 mc. The noise is most severe at the frequency where the device producing it is large enough to act as a good antenna. For example, vehicles whose dimensions approach a half-wavelength in the vicinity of 30 mc produce their most severe ignition noise in this region. Most equipment used for vehicular communication, as well as some important fixed stations, operates in this range. The noise produced by the discharge of current through a gas, mercury-vapor rectifiers, and similar devices has a spectrum that extends up to extremely high frequencies with very high intensities. It often is impossible to remove the noise from these devices because it is impractical to shield them.

e. Natural Noises. Natural noises disturbing to radio communication arise from various sources and may be either impulse or fluctuation types. Perhaps the most familiar are the frequent and overlapping noise in pulses produced by lightning discharges. This noise originates not only in local storms but also in the tropical storm centers, from which it is propagated as a radio wave to many parts of the earth. The signal strength of noise produced by local storms decreases directly with increasing frequency, the frequencies above 40 mc being less subject to this interference than the lower frequencies. The storms are more intense in summer than in winter, and, since most of the static occurs in the warm temperatures, the lowest noise levels are found in the colder weather. There are also weaker noise impulses thought to be produced by sources not on the earth, such as the noise attributable to sun spots. Naturally, nothing can be done to suppress natural noises at their sources, whether they originate near the receiver or not. Therefore it is desirable to have some other means of overcoming them.

f. Receiver Noise. The limiting factor governing the sensitivity of most high-frequency receivers is the amount of noise inherent in the receiver. Although there may be no defective parts, there still will be a quantity of noise in even the best receivers. This noise is of a random fluctuation type, producing a characteristic hiss-like sound in the loudspeaker or earphones. The most important noise in very-high-frequency equipment, where noise outside the receiver is low enough to permit receiver noise to be noticed, is caused by the current flowing through resistors and tubes.

g. Communication in Presence of Noise. A general communication system is shown in the block diagram of figure 33. The information source supplies a message which is converted into an electrical impulse. It then is trans-

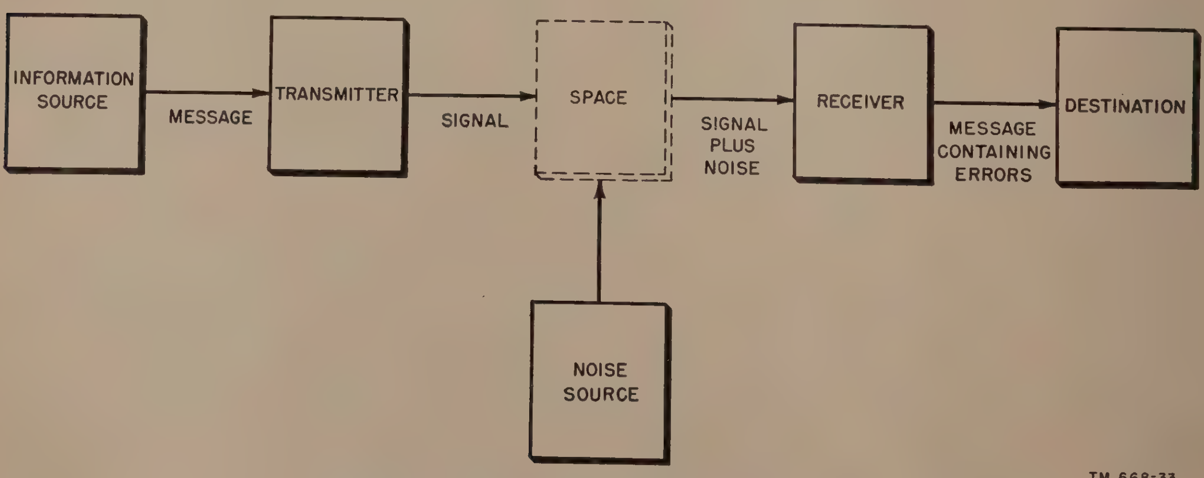


Figure 33. Noise in general communication system.

formed into a signal that can be sent through space to the receiver. The receiver recovers the signal and reproduces the message. During the course of the transmission, the signal may be modified by the addition of some noise, and therefore the receiver cannot reconstruct the original message perfectly. However, in an f-m system, if the deviation is increased above a certain minimum signal-to-noise ratio, the message-to-error (or degree-of-accuracy) ratio increases.

h. Noise in F-M Reception. The effect of random fluctuation noise and impulse noise in f-m receivers differs. Fluctuation noise pulses excite tuned circuits in the receiver, causing them to oscillate at their resonant frequency. The oscillations interfere with the carrier, causing spurious noise to appear in the detector output, and then diminish gradually until the next noise pulse. Impulse noise is largely a sudden disturbance in the amplitude of the signal. It is removed by the f-m detector, which does not respond to the amplitude variations of the carrier. If the frequency-modulated signal is very weak compared with the noise, the intelligence-containing side bands are suppressed in the f-m detector. Therefore, if the carrier is not above a certain minimum amount, f-m is actually worse than a-m. The value of this minimum is called the *threshold of improvement* and depends on the frequency deviation. Only when the signal is above this threshold is wide-band f-m superior to a-m. Narrow-band f-m is as good as a-m at low signal levels and is equal or inferior to it at high levels.

i. F-M and A-M Response to Impulse and Fluctuation Noise. In figure 34 the signal-to-noise ratio is plotted vertically against the peak carrier-to-noise ratio. Amplitude modulation is shown by the straight diagonal line for comparison purposes. For a modulation index of 1, the dashed lines indicate the improvement in signal-to-noise ratio over a-m. The lower of the pair, labeled F_1, is the improvement for fluctuation noise, and the higher, I_1, is the improvement for impulse noise. The other set of dotted lines shows the same situation for a modulation index of 4, with curve F_4 for fluctuation noise and I_4 for impulse noise. The threshold of improvement begins where the carrier is sufficiently strong that the curve crosses the reference a-m line.

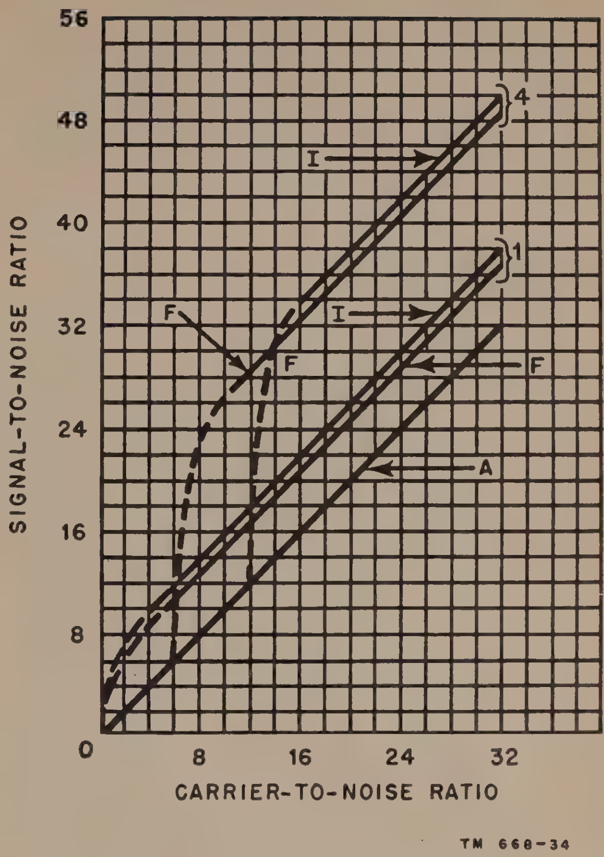


Figure 34. F-m noise characteristics.

j. Noise in Presence of Modulation. These analyses have been made with sine-wave modulation. With speech or other forms of information impressed on the carrier, the results are somewhat different, but in general they agree with that of the sine-wave. For normal military communication, f-m is an improvement over a-m because of its higher immunity to noise. *Narrow-band* f-m is preferred where the carrier frequency is relatively low and channel space is at a premium. *Wide-band* f-m is used at the higher frequencies, where this is not a problem. However, when the signal is weak, wide-band f-m is inferior in performance to an equivalent a-m system. In this it differs from narrow-band f-m, which is capable of excellent performance at low-signal levels. The reduction in errors caused by poor signal-to-noise ratio also makes advantageous the use of f-m in radiotelegraphy systems. In these systems, the sig-

nal is modulated by either a series of rectangular pulses or the use of two continuous audio tones as modulating signals to shift from one tone to the other as the transmitter is keyed. The latter method has the advantage of requiring somewhat less spectrum space than the rectangular modulation method.

15. Interference

a. General.

(1) Any receiver of radio signals, no matter what type of modulation is used, has a certain bandwidth over which it will accept signals. In this *acceptance band,* the receiver reproduces signals at the output of the detector for which it was designed. An a-m receiver must be capable of passing all frequencies that go into the modulation side bands on either side of the carrier in addition to the carrier itself. If speech is to be received, the receiver must have a bandwidth at least twice the highest speech frequency that is transmitted. For good voice transmission the highest audio frequency is usually 3 kc. Therefore, the bandwidth of an a-m receiver for the reception of speech must have an acceptance band of at least 6 kc.

(2) An f-m receiver must have a bandwidth that is at least enough to allow for the full frequency deviation plus the side bands of the transmitted signal. In a receiver with a given bandwidth, any spurious or unwanted signals that fall within the acceptance band give rise to some form of interference to the desired signal. In addition, no practical receiver is made with an acceptance band that just covers the necessary bandwidth with no response outside those limits. If the signal that interferes with the desired signal is in the same channel, the interference is called *co-channel interference.* If the disturbance originates on either side of the receiver acceptance band, it usually is called *adjacent channel interference.* The receiver also can have *spurious responses* at frequencies other than the one at which it is tuned because of inadequacies in the rejection of unwanted frequencies in the first receiver stage.

b. A-M.

(1) Within the acceptance band of the a-m receiver, any unwanted side bands are equivalent to spurious modulation, and they are amplified and detected just as the desired signal is. Therefore, no matter what steps are taken in the design of the receiver, these undesired signals cause an output to appear in the loudspeaker or earphones. The behavior of a receiver in the presence of interference can be analyzed in connection with the over-all acceptance band of the receiver, and the results for a-m are different from those for f-m.

(2) To see what happens when two signals combine in the pass band of any type of receiver, consider the vector diagram of figure 35. In this diagram, the undesired carrier is represented by the vector labeled E_i. It is not a stationary vector but is generating a sine wave and therefore is rotating at a definite rate depending on the frequency. Assume that it combines with the desired carrier, E_c, which has a frequency that is slightly different from that of the undesired carrier, E_i. Although both carriers are rotating, it is assumed that E_c is standing still while E_i rotates at a rate equal to the difference between its natural frequency and that of the desired carrier. Since the two carriers are added in the receiver, the resultant vector shown by the arrow, E_r, has a frequency equal to the difference between the two carriers. Moreover, its relative phase, a, is continuously changing with the rotation of the carriers, so that the resultant is phase-modulated. Since the frequency difference between the two signals was assumed to be small, the vector changes in length as the relative angles of its components

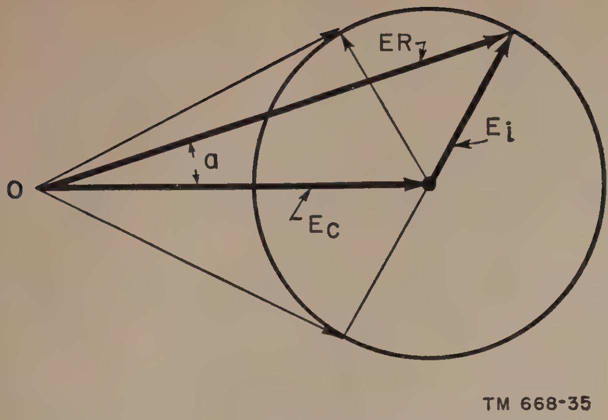


Figure 35. Interference between two unmodulated carriers.

vary. Therefore, it is also amplitude-modulated at the difference frequency.

(3) The single interference resultant produced between two unmodulated carriers is modulated in both amplitude and phase. Since frequency and phase modulation differ so little, it can be considered modulated also in frequency. This composite modulation is called a *beat note* between the two carriers. When this beat note is received, it produces spurious signals that vary both in frequency and in amplitude and therefore appear in the output of the receiver, regardless of the kind of modulation for which the receiver is designed. The output of the detector also contains simple multiples of the fundamental beat note. These are its harmonics and they are strong when the two carriers are nearly equal in amplitude.

(4) When one of the carriers is amplitude-modulated, the resultant interference varies at the modulating frequency. The beat note here too is frequency- , phase- , and amplitude-modulated, and can interfere in any receiver. If both carriers are amplitude-modulated, the resultant beat note is heavily amplitude-modulated, even though the strength of the interfering carrier is a fraction of the strength of the desired one. In fact, it can be shown by mathematical analysis that the interfering signal can be as much as one thousand times weaker than the desired carrier and still cause an audible beat note. If it is more than one-hundredth as strong, it causes prohibitive interference. Therefore, it can be seen that for military purposes, where there must be many pieces of equipment in use at once, interference-free reception with a-m may be impossible under emergency conditions.

c. F-M. In f-m, modulation of an otherwise unmodulated carrier by the undesired signal decreases as the deviation of the undesired carrier increases. In fact, it never can exceed one-half cycle, which is equivalent to a modulation index of one-half. When the desired carrier is modulated, the interference decreases as the deviation is increased, especially since the f-m detector is assumed to be insensitive to a-m. Wide-band f-m is less susceptible to interference than a-m, and narrow-band f-m gives some improvement. If the two interfering signals differ in amplitude by a relatively small ratio of two to one, the interference between the carriers is negligible. Therefore, because of its greater freedom from interference, f-m is more widely used for military applications than a-m. In f-m, the stronger of two wide-band, or even two narrow-band, transmitters is received practically without interference from the other.

16. Summary

a. In a frequency-modulated wave, the instantaneous frequency varies about the carrier frequency in proportion to the amplitude of the modulating signal.

b. The variations in instantaneous frequency are determined by the frequency of the modulating wave; the higher the modulating frequency, the greater the number of deviations in a given time period.

c. In amplitude modulation, two side-band frequencies (an upper and a lower) are generated for each sine-wave component of the modulating signal.

d. In an f-m system, many side bands are generated for each sine-wave component of the modulating signal.

e. The bandwidth occupied by an f-m signal exceeds the peak deviation limits. The total bandwidth increases for increasing modulation frequency, all other things being constant.

f. When the channel space occupied by an f-m transmitter approximates that of an a-m transmitter for the same modulating signal, the transmission is termed narrow-band f-m.

g. Where f-m bandwidth greatly exceeds that of the equivalent a-m signal, it is called wide-band f-m.

h. The amplitudes of the side bands, as well as their frequencies, depend on the amplitude and frequency of the modulating signal and the frequency deviation of the transmitter.

i. For a sinusoidally modulated signal, the side bands are distributed in symmetrical pairs on either side of the carrier frequency at integral multiples of the modulation frequency.

j. The ratio of the frequency deviation to the frequency of the modulating signal is called the modulation index.

k. The value of the modulation index determines the amplitude and number of the various side bands.

l. The number of side-band pairs increases as the modulating frequency decreases.

m. To allow for the side bands beyond the peak deviation limits, additional channel space, called a guard band, is provided on either side of the channel.

n. For nonsinusoidal modulation, the side-band frequency distribution depends on the sinusoidal components present in the nonsinusoidal wave, and can be analyzed in terms of each component separately plus all possible products of each side band with every other. The bandwidth for a nonsinusoidally modulated signal generally is greater than that for a sine wave of the same amplitude and frequency.

o. Where the modulation is in the form of rectangular waves, the result is termed pulse modulation. The number of pulses per second is called the pulse repetition rate.

p. Symmetrical modulating signals produce a symmetrical side-band spectrum.

q. Because the high frequencies in a sound signal are weaker than the low ones, they are amplified to a greater extent in many transmitters to improve the ratio of signal to noise in this range. This is called preemphasis. The reverse process is called deemphasis and is provided for in the receiver.

r. Radio noise is of two types—impulse, or sharp r-f pulse voltage; and continuous, or fluctuating, signals of random amplitude and phase.

s. Improvement in the signal-to-noise ratio takes place only with a minimum value of carrier present at a level termed the threshold of improvement.

t. Frequency modulation is less susceptible to interference than a-m because the maximum effective modulation index of the interfering wave can never exceed one-half if the desired carrier is twice as strong as the undesired carrier.

17. Review Questions

a. What happens to the frequency deviation of an f-m wave when the amplitude of the modulating signal is increased?

b. If the frequency of the modulating signal is doubled, what happens to the f-m wave? Describe in detail.

c. What determines the effective bandwidth of an a-m transmitter?

d. Why are guard bands used between adjacent f-m channels?

e. Are the side bands of an f-m transmitter the same for sinusoidal and nonsinusoidal modulating voltages?

f. How is it possible to use a larger bandwidth for an f-m transmission than for an a-m transmission?

g. When does f-m have only two side bands?

h. Which f-m signal has fewer side bands, one with a higher or one with a lower modulating frequency, for the same total deviation?

i. What is the minimum bandwidth of a narrow-band f-m signal?

j. If the audio-modulating frequency is the same for two f-m transmitters, but one is adjusted for one-half the deviation of the other, which will have the greatest effective number of side-band pairs? Why?

k. If a 5-kc audio signal produces a deviation of 15 kc at 33 mc, what is the maximum instantaneous frequency of the f-m wave?

l. What is meant by the modulation index?

m. What is the modulation index for an audio signal of 2 kc with a deviation of 220 kc? What total bandwidth would such a signal occupy?

n. Define pulse repetition rate.

o. What is the effect on the side-band spectrum when two signals of different amplitude and frequency simultaneously modulate the carrier?

p. When is the spectrum that results from nonsinusoidal modulation symmetrical? When is it not symmetrical?

q. What are the highest and the lowest frequencies that need to be transmitted for intelligible speech?

r. What is the difference in the relationship of phase-to-frequency deviation between a direct f-m transmitter and an indirect f-m transmitter, modulated by speech?

s. What is the purpose of preemphasis?

t. What is the time constant of a deemphasis network consisting of a capacitor of .003 microfarad in series with a resistor of .1 megohm?

u. What is the difference between impulse and fluctuation noise?

v. In an f-m communication system, what is the relationship between the bandwidth and noise?

w. By how much must two carriers differ in amplitude if they are not to interfere with each other when they are both amplitude-modulated?

x. Which is better for the reduction of interference for stations operating on the same channel—narrow- or wide-band f-m? Why?

CHAPTER 3

METHODS OF PRODUCING FREQUENCY MODULATION

Section I. DIRECT METHODS OF PRODUCING FREQUENCY MODULATION

18. General

a. Methods for producing f-m are *direct* and *indirect*. In the former, the change in frequency of the oscillator is directly proportional to the amplitude of the audio signal. With indirect methods, the change in phase angle is directly proportional to the amplitude of the audio signal.

b. The two basic transmitter stages necessary for producing f-m are the r-f oscillator and the modulator. The oscillator can be any one of the basic types, such as the Hartley or the Colpitts. The modulator stage provides a way of controlling oscillator frequency, usually with a modulating signal, at some audio frequency. A discussion of methods for producing f-m therefore is essentially a discussion of different modulator circuits.

c. Figure 36 shows a simple shunt-fed Hartley oscillator. Feedback is accomplished by coupling energy from the plate to the grid circuit through the split inductance. This feedback sustains oscillations at a resonant frequency determined mainly by the values of inductance and capacitance in the tank circuit. The frequency can be determined approximately by means of the formula,

$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

where

f_o is the resonant frequency
L is the inductance
C is the capacitance

From this formula, it can be seen that a change in either inductance or capacitance results in a change in the resonant frequency. If the inductance or capacitance is increased, the resonant frequency is lowered; if L or C is decreased, f_o is raised. Since the frequency can be changed by changing L or C, the next step is to find a method for allowing the audio signal to control either L or C.

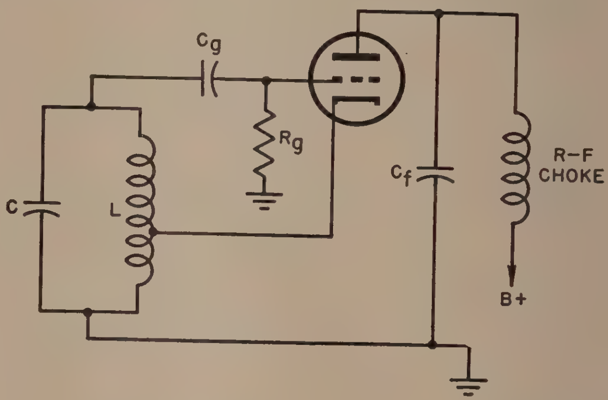


Figure 36. Shunt-fed Hartley oscillator.

19. Mechanical Methods

a. Changing Capacitance. In figure 37, a capacitor-type microphone is substituted for C in the tank circuit of the oscillator. The two plates of the microphone, one fixed and one movable, have some value of capacitance between them. This capacitance, C, together with the inductance, L, forms a tank circuit resonant at some frequency, f_o. When sound waves strike the movable plate (diaphragm) of the microphone, it moves back and forth with the changes in air pressure caused by the sound. From basic theory, it is known that changing the distance, d, between the plates of a capacitor changes its capacitance; the greater the distance, the

smaller the capacitance; the smaller the distance, the greater the capacitance. As the diaphragm moves back and forth, the capacitance of the tank circuit is being changed accordingly. A changing capacitance in the tank circuit means a changing resonant frequency. Since the audio signal controls the movement of the diaphragm and the value of capacitance, it also determines the resonant frequency. The output is a frequency-modulated wave, whose frequency deviation is controlled by the amplitude of the audio signal. There are certain drawbacks to using the circuit just described to produce an f-m wave since only a capacitor-type microphone can be used. This restriction is undesirable in view of the many types of microphones in use. Furthermore, it is necessary, at times, to separate the microphone from the transmitter by considerable distances. This arrangement is impossible with the circuit of figure 37.

specially designed signal generators. However, the mechanical system necessary is not capable of changing frequency in accordance with voice-modulating frequencies. For this reason, it cannot be used in f-m transmitters.

20. Reactance

Since the reactance of a capacitor, X_c, is equal to

$$X_c = \frac{1}{2\pi f_c} \text{ ohms}$$

Where

X_c is capacitive reactance

f is the frequency of the voltage source

C is the capacitance

It can be seen that, if the capacitance is held constant, the frequency must be changed in order to change the reactance. The same effect

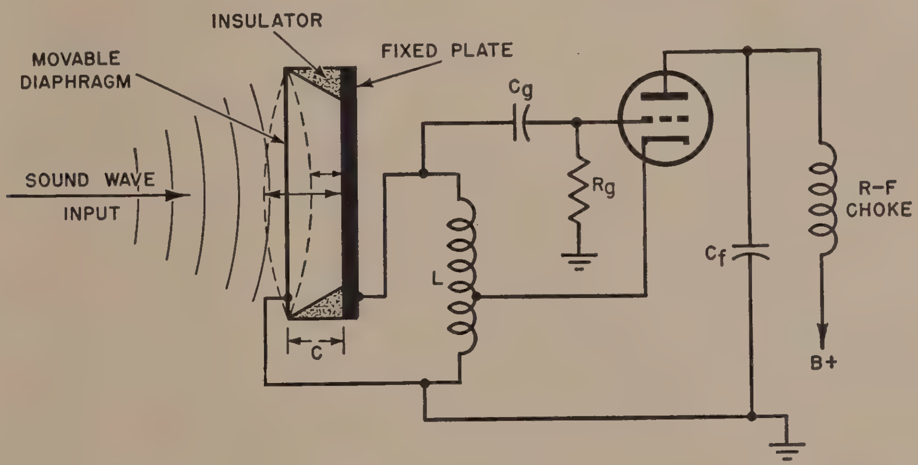


Figure 37. Capacitor-type microphone produces f-m.

b. Changing Inductance. The inductance of a coil constructed with a powdered-iron slug at its center can be varied by moving the slug in or out of the coil. If this coil is substituted for L in the tank circuit of the Hartley oscillator, the resonant frequency can be made to depend upon the position of the slug in the coil. By using a motor and a suitable gear train, the slug can be made to move in and out of the coil at a rate determined by the motor speed, and the resonant frequency is changed accordingly. This system is used to produce an f-m wave in

can be noted by inspecting the formula for inductive reactance:

$$X_L = 2\pi f L \text{ ohms,}$$

where

X_L is the inductive reactance

f is the frequency

L is the inductance

It is possible to control the frequency of an oscillator by controlling the amount of reactance, capacitive or inductive, which is present in the circuit. This is accomplished in a circuit known as a *reactance modulator,* which uses

the characteristics of a vacuum tube to control the reactance in the tank circuit of the oscillator. By simulating a capacitance or an inductance across its output terminals, it is said to *inject* reactance into the tank circuit. The simulated capacitance or inductance in turn is controlled by the audio signal.

21. Vacuum-Tube Characteristics

a. Transconductance. Before proceeding with the discussion of reactance modulators, it is desirable to review briefly the vacuum-tube characteristics on which circuit operation depends. The transconductance of a vacuum tube is defined as the ratio of a small change in plate current to the small change in the grid voltage that produced it, with the plate voltage held constant. This ratio can be expressed mathematically in the form:

$$g_m = \frac{di_b}{de_c}$$

where

g_m is the transconductance
i_b is the plate current
e_c is the grid voltage
d signifies *a small change in*

Transconductance is a form of conductance, and it is measured in *mhos*, where the current is in amperes and the voltage in volts. Because the transconductance almost never exceeds fractional values, a smaller unit, the *micromho*, one-millionth of a mho, is used. The transconductance of a vacuum tube under rated operating conditions usually is listed in tube manuals.

b. Amplification Factor. The amplification factor, μ, or mu, of a vacuum tube is defined as the ratio of a small change in plate voltage to the small change in grid voltage that produced it, the plate current being held constant. Expressed mathematically:

$$\mu = \frac{de_b}{de_c}$$

where

μ is the amplification factor
de_b is the change in plate voltage,
de_c is the change in grid voltage.

The μ of a tube determines how effective the grid is in controlling the plate current. For example, if the μ is 35, it means that the grid voltage is 35 times more effective in controlling the plate current than the plate voltage. For a given change in plate current, therefore, the change in plate voltage would have to be 35 times the necessary change in grid voltage.

c. Plate Resistance. The plate resistance of a vacuum tube is defined as the ratio of a small change in plate voltage to the small change in plate current that produced it, the grid voltage being held constant. In mathematical form—

$$r_p = \frac{de_b}{di_b}$$

where

r_p is the a-c resistance from plate to cathode
de_b is a small change in plate voltage
di_b is a small change in plate current

d. Relationship of g_m, μ, and r_p. It is possible to combine the three equations for transconductance, amplification factor, and plate resistance so that a direct relationship exists between them. If the equations for g_m and r_p are multiplied

$$g_m \times r_p = \frac{di_b}{de_c} \times \frac{de_b}{di_b}$$

The di_b terms cancel out, and

$$g_m \times r_p = \frac{de_b}{de_c}$$

An inspection of the equation shows that the right-hand side is the expression for the amplification factor, μ. Therefore, μ can be substituted for de_b/de_c, and

$$g_m \times r_p = \mu$$

The significance of this relationship between vacuum-tube characteristics will become apparent in considering reactance modulator circuits as used in f-m transmitters.

22. Reactance-Tube Modulator

The purpose of the reactance-tube modulator in a direct f-m transmitter is to frequency-modulate the f-m signal of an oscillator in accordance with the audio signal. The modulator accomplishes this by changing the amplitude variations of the audio signal into a varying reactance which is injected into the tank circuit of the oscillator. This injected reactance changes the frequency of the oscillator.

a. Basic Circuit.

(1) The fundamental arrangement of the circuit of a reactance-tube modulator and oscillator is shown in figure 38. At the extreme left, the audio signal is impressed between terminals 1 and

2, or between grid and ground of the reactance tube. Therefore, the audio signal controls the reactance-tube plate current. The plate current flows through a load consisting of Z_a and Z_b in series, which is connected across the tank circuit of the oscillator through terminals 3 and 4. When the plate current is varied by the audio signal, the reactance of the plate load is changed, thus changing the operating frequency of the oscillator tank circuit. The resultant f-m signal then is coupled inductively to the following stages of the transmitter through terminals 5 and 6.

(2) With no audio signal present at the grid of the reactance tube, its plate current has some small d-c value whose effect on the plate load is negligible. The r-f voltage present across the oscillator tank circuit also is impressed across the reactance tube plate load. The plate load of Z_a and Z_b presents a reactance to this voltage which can simulate the action of either a capacitor or an inductor. This is the same as adding a capacitor or an inductor in parallel with the tank circuit, which changes either the effective capacitance or the inductance of the tank circuit. The operating frequency of the oscillator, therefore, depends not only on L and C, but also on Z_a and Z_b. Since this operating frequency is the oscillator output frequency with no audio signal present, it also must be the center frequency of the f-m signal. The audio signal, when applied between terminals 1 and 2, is impressed not only from grid to ground, but also across Z_b in the plate circuit. The combined effect of these two actions is to change the reactance of Z_a and Z_b in accordance with the audio signal.

b. Impedance at Terminals 3 and 4.

(1) Looking from the tank circuit toward terminals 3 and 4, an impedance caused by Z_a, Z_b, and the reactance tube is seen. For a given amplitude of audio signal, the resultant plate current

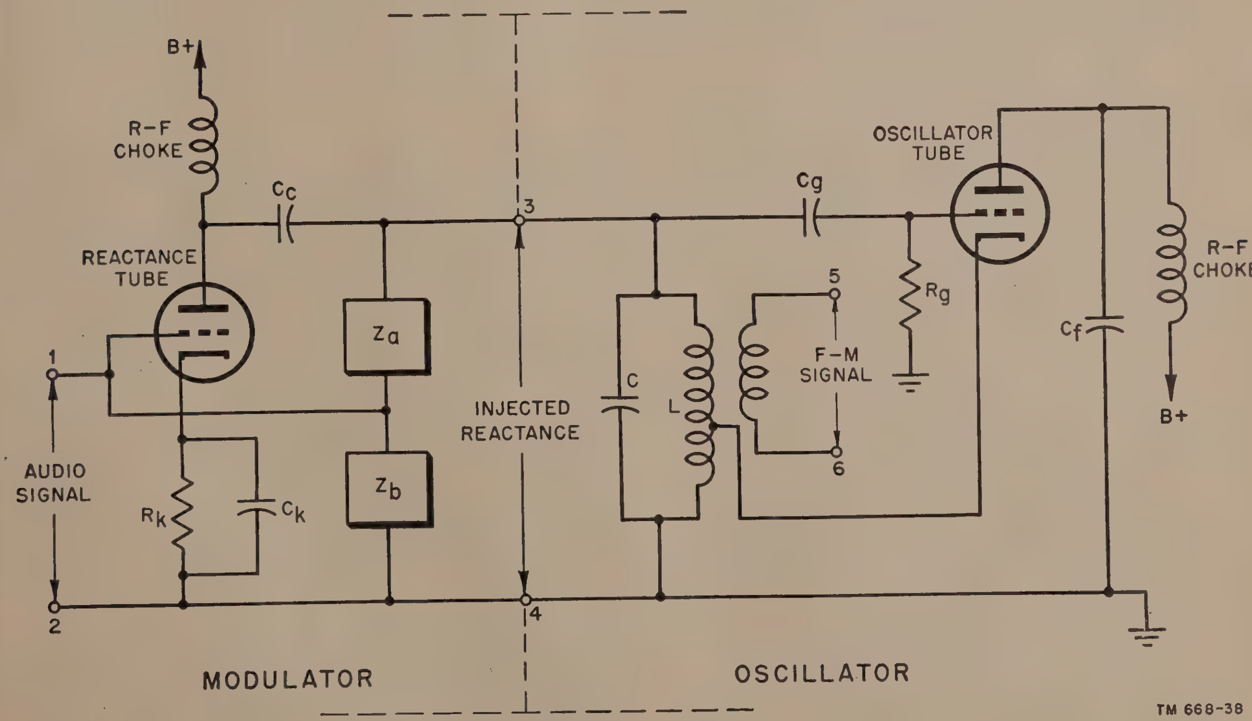


Figure 38. Basic circuit of reactance-tube modulator and oscillator.

through the tube can be found from from the tube transconductance

$$g_m = \frac{di_b}{de_c}$$

or

$$di_b = g_m \times de_c$$

The plate current flowing through the plate load is

$$i_b = g_m \times e_c$$

where the plate load consists of Z_a and Z_b, or Z_{ab}. In circuits of this type, the plate resistance of the tube usually is high enough to be ignored when considering the plate load.

(2) If the voltage across Z_{ab} is E_{ab}, then (by Ohm's law):

$$Z_{ab} = \frac{E_{ab}}{i_b}$$

Substituting for i_b,

$$Z_{ab} = \frac{E_{ab}}{g_m \times e_c}$$

From this it can be seen that the transconductance of the tube has an inverse effect on impedance Z_{ab} across the oscillator tank circuit. It remains now to break e_c into its equivalent components in order to observe all the factors at work in the determination of this impedance.

(3) An examination of figure 38 shows that e_c is impressed across Z_b in the plate circuit of the reactance tube. If the total voltage across Z_{ab} is E_{ab}, then e_c must be related to E_{ab} as Z_b is related to Z_{ab}. In mathematical form, this relation can be expressed as:

$$\frac{e_c}{E_{ab}} = \frac{Z_b}{Z_a + Z_b}$$

When this equation is solved for e_c, it is found that

$$e_c = E_{ab} \times \frac{Z_b}{Z_a + Z_b}$$

This value of e_c then is substituted in the equation for the total impedance found in the previous paragraph:

$$Z_{ab} = \frac{E_{ab}}{g_m \times e_c}$$

$$Z_{ab} = \frac{E_{ab}}{g_m \times E_{ab} \times \frac{Z_b}{Z_a + Z_b}}$$

The E_{ab} terms cancel out, and the remaining terms of the equation can be simplified in the following manner:

$$Z_{ab} = \frac{1}{g_m} \times \frac{Z_a + Z_b}{Z_b}$$

$$= \frac{1}{g_m} \times \frac{(1 + Z_a)}{Z_b}$$

$$= \frac{1}{g_m} + \frac{1}{g_m} \times \frac{Z_a}{Z_b}$$

The last form of the equation gives the impedance at terminals 3 and 4 in terms of the reactance-tube transconductance and the impedance network of Z_a and Z_b.

(4) This equation for the impedance seen by the oscillator tank circuit expresses mathematically the principle of operation of the reactance-tube modulator. The impedance, Z_{ab}, is the general term for the total impedance across the oscillator tank circuit. The terms Z_a and Z_b are expressions for the circuit components which together constitute the plate load. These can be capacitors, inductors, or resistors which have *fixed values* in the circuit, and the only way to vary Z_{ab} is to vary the transconductance of the reactance tube. This takes place when an audio signal is applied to the input to the modulator.

c. Injected Reactance. An examination of the equation for the total impedance shows that it contains two parts. The first part is the reciprocal of a transconductance, which is resistance. The second part contains two terms for impedance, Z_a and Z_b. If either Z_a *or* Z_b is made reactive, the second part of the equation, $1/g_m$ times Z_a/Z_b, is a form of reactance. This reactive component is the injected reactance. Assume that Z_a is a fixed capacitor having a reactance of X_c at a frequency, f. If Z_b is a resistor having a resistance, R, then the injected reactance, X_i, *is*

$$X_i = \frac{1}{g_m} \times \frac{X_c}{R}$$

d. Circuit Operation. Figure 39 shows a reactance-tube modulator circuit with C_L and R_L forming the plate load. With no audio signal present, this network injects a fixed reactance across the tank circuit which determines the

operating frequency of the oscillator or center frequency of the f-m wave. When an audio signal appears at the input, the injected reactance is varied above and below its zero-signal value, thereby varying the frequency of the oscillator above and below the center frequency. The variation in injected reactance, as well as the frequency deviation, depends on the amplitude of the modulating signal.

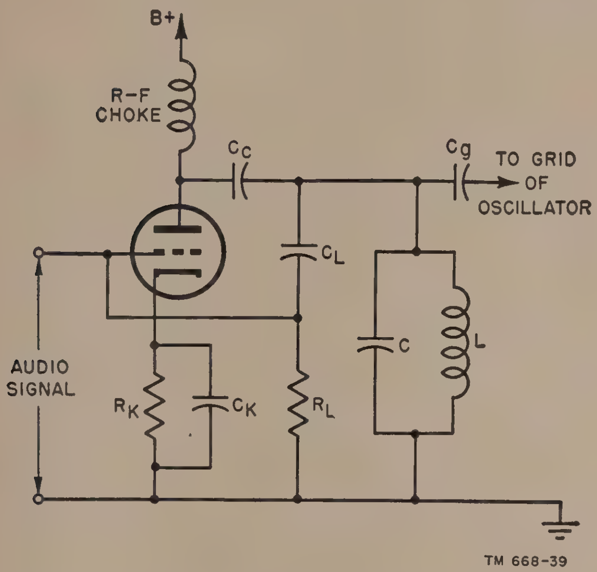


Figure 39. Typical reactance-tube modulator circuit.

(1) *Zero-signal operation.*

(*a*) With no audio signal present, the only signal impressed on the modulator circuit is that of the r-f voltage across the oscillator grid circuit. This voltage is applied across the plate load of the modulator, or across C_L and R_L. The reactance of C_L is made very large in respect to the resistance of R_L, and therefore the capacitive reactance, XC_L, determines the resultant current flow, causing it to lead the voltage across it by approximately 90°. This current flowing through R_L results in a voltage drop across R_L which also leads the applied voltage by 90°. Actually, the voltage across the resistance is in phase with the current through it; however, it can be seen that the voltage across R_L is applied between grid and ground of the reactance tube.

(*b*) This r-f voltage at the grid of the reactance tube causes an r-f variation of plate current which is coupled back to the tank circuit of the oscillator through the coupling capacitor, C_c. However, the current in the oscillator tank circuit is in phase with the r-f voltage, since the circuit is operating at resonance, while this additional current resulting from the *same* voltage leads the voltage by 90°. The additional current supplied by the reactance tube *acts as if it were caused by a capacitor.*

(*c*) The circuit operation just described can be seen in terms of vectors in figure 40. In A, the tank circuit voltage, E_T, and the current, I_T, are in phase because the circuit is at resonance. The r-f voltage of the tank circuit, E_T, causes a current flow through the plate load of the reactance tube. This current, i_L, leads the applied voltage by 90°, because the capacitive reactance of C_L has been made large in comparison with the resistance of R_L. In B, this current, i_L, flows through R_L, causing a voltage drop, E_{RL}, across the resistor which is in phase with the current through it. The circuit, however, is arranged so that this voltage is coupled to the grid of the reactance tube; E_{RL} is also e_c. In C, this grid voltage causes an r-f variation in plate current, i_b, which is in phase with the grid voltage. The capacitor, C_c, has negligible reactance and when the r-f current is coupled to the oscillator tank, it adds to the current flowing in it. Since i_b is 90° out of phase with the tank current and leading it, i_b acts as if it were coming from a circuit component having capacitive reactance, and consequently a capacitance.

(2) *Zero-signal injected reactance.* The amount of reactance injected in this

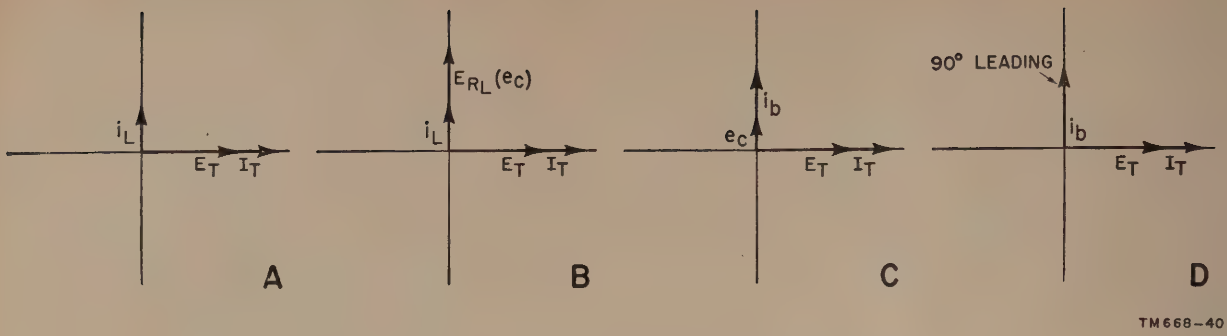


Figure 40. Vector representation of zero-signal operation.

circuit can be calculated from the formula

$$X_i = \frac{1}{g_m} \times \frac{X_{CL}}{R_L}$$

where

X_i is the injected reactance
g_m is the reactance-tube transconductance
X_{CL} is the capacitive reactance
R_L is the resistance

From the analysis of this particular arrangement of the plate load components, it is known that X_i is a capacitive reactance. Basic theory shows that the capacitive reactance of any capacitor can be found from the formula

$$X_c = \frac{1}{2\pi fC}$$

Since this is true, similar expressions can be written for the capacitive reactance of C_L and also for C_i,

$$X_c = \frac{1}{2\pi fC_L}$$

and

$$X_i = \frac{1}{2\pi fC_i}$$

Substituting these expressions for X_{CL} and X_i in the formula for the injected reactance:

$$\frac{1}{2\pi fC_i} = \frac{\frac{1}{2\pi fC_L}}{R} \times \frac{1}{g_m}$$

When this expression is simplified, the injected capacitance, C_i, is found to be

$$C_i = g_m \times R_L \times C_L$$

The effect of the reactance modulator, with no audio signal present, is the same as that of a capacitor having C_i capacitance placed across the oscillator tank. Since C_i is in parallel with C of the tank circuit, the two capacitances add, increasing the total capacitance and decreasing the operating frequency some fixed amount. The new operating frequency then becomes the center frequency of the f-m wave.

(3) *Effect of audio signal.* Applying an audio signal at some single frequency to the grid of the reactance tube causes two voltages to be present at the grid—an a-f voltage and an r-f voltage. The r-f voltage is responsible for the reactive plate current flow and the audio signal *changes* the amount of plate current flowing in accordance with its amplitude. Changing the amount of plate current coupled to the tank circuit means that its reactive effect is varied and results in the injection of a changing reactance into the oscillator tank. This changing reactance adds a changing capacitance to the oscillator tank. Oscillator frequency is varied accordingly and the result is a frequency-modulated signal at the oscillator output. The effect of the audio signal can be seen in terms of the formula for injected capacitance, where C_i is equal to $g_m R_L C_L$. The variations in a-f voltage at the grid have the same effect on the r-f plate current as a varying tube transconductance. From the formula above, it can be seen that varying the transconductance changes the injected capacitance, and causes frequency modulation to appear at the oscillator output.

e. Plate-Load Arrangements.

(1) The arrangement shown in figure 39 is not the only possible arrangement that can be used to inject a reactance into the oscillator. For example, the capacitor and resistor used in the basic circuit can be reversed so that *inductive* reactance is injected into the oscillator. A combination of an inductor and a resistor or any combination but that of capacitor and inductor allows the circuit to frequency-modulate the oscillator output. Four arrangements are possible: capacitor-resistor, resistor-capacitor, inductor-resistor, and resistor-inductor. However, the formula for the injected reactance involving g_m, Z_a, and Z_b has a different arrangement for each circuit.

(2) In figure 41, a resistor is substituted for Z_a and a capacitor for Z_b. The components are chosen so that the resistance of R_L is much greater than the reactance of C_L. Since the resistive component is so much larger, the r-f voltage applied to the plate load by the tank circuit causes the current to be in phase with the r-f voltage. As in any capacitor, however, the current leads the voltage by 90°, and this is true also of C_L. The voltage across C_L, therefore, lags the current and the applied voltage by 90°. This voltage across C_L is coupled to the grid of the reactance tube, and causes an r-f variation in plate current that is in phase with the grid voltage. The r-f current is coupled to the oscillator tank, and, since it is in phase with the grid voltage, it must lag the current in the tank by 90°. This produces the same result as when an *inductor* is placed across the tank. The modulator therefore is said to inject inductive reactance into the tank circuit. The amount of simulated inductance necessary to produce this injected inductive reactance is:

$$L_i = \frac{R_L C_L}{g_m}$$

(3) Keeping the large resistor, R_L, for Z_a, it is possible to substitute a small inductor for Z_b (fig. 42). With this arrangement, the circuit now injects a *capacitive* reactance into the oscillator tank circuit. The oscillator voltage applied across the plate load of the reactance tube causes a current to flow whose phase is controlled by the large resistance of R_L. This current is in phase with the applied voltage, since R_L is large in respect to X_{LL}. The voltage across any inductor, however, leads the current through it by 90°. The voltage across the inductor, therefore, leads both the current and the applied voltage by that amount. Because this voltage is coupled to the grid of the reactance tube, an r-f plate current flows which is in phase with the grid voltage and 90° leading in respect to the oscillator tank voltage. When coupled to the oscillator tank, this current acts as if it were caused by a *capacitor* having a capacitive reactance equal to the injected reactance. The amount of this capacitance can be found from the formula,

$$C_i = \frac{g_m L_L}{R_L}$$

(4) When the resistor-inductor arrangement is reversed (fig. 43), the circuit injects *inductive* reactance into the oscillator tank. The inductance of L_L is

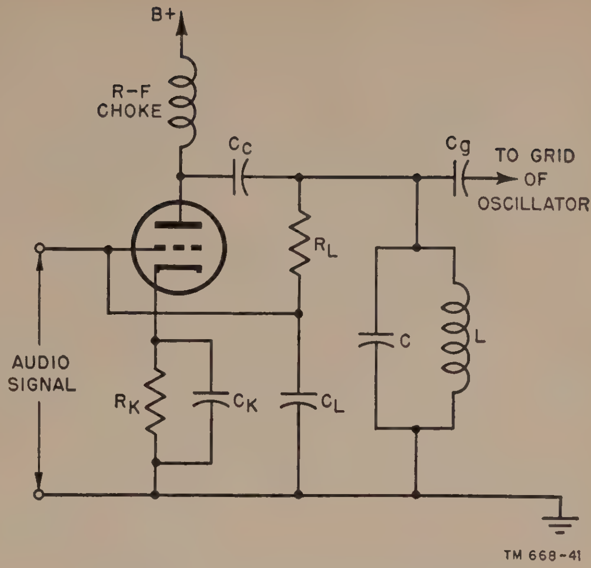


Figure 41. R-c reactance-tube load.

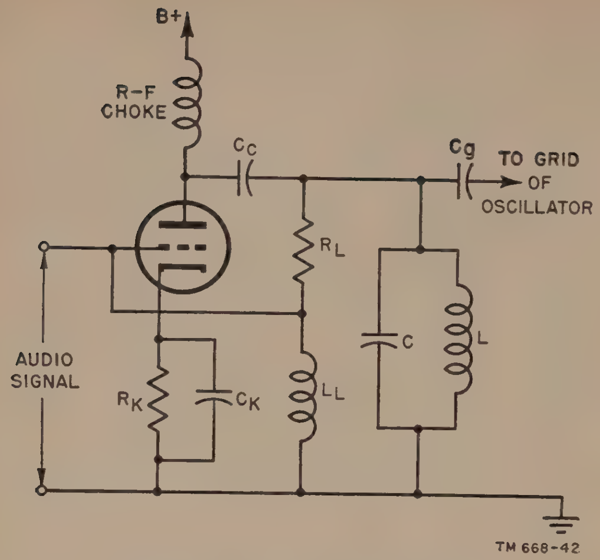


Figure 42. R-L reactance-tube load.

made large to increase the inductive reactance in respect to R_L. The r-f voltage from the oscillator tank then will cause a current to flow through the plate load which lags the applied voltage by 90°. This voltage then is applied to the grid of the reactance tube, producing an r-f plate current which is lagging the current in the tank circuit by 90°. Since a lagging current is an indication of inductance, the modulator simulates the action of an inductor placed across the tank circuit. The amount of injected inductance can be found from the formula

$$L_i = \frac{L_L}{g_m R_L}$$

(5) The four arrangements possible in the plate circuit of the reactance tube are shown in figure 44, together with the formulas for calculating the injected inductance or capacitance. It must be borne in mind that Z_a always is made large in respect to Z_b, and it is this component which determines the phase of the current caused by the r-f oscillator voltage. The amplitude of the reactive current is controlled by the amplitude of the audio-modulating signal by means of the tube transconductance. Z_a and Z_b, therefore, control the operating frequency of the oscillator, and g_m controls the frequency deviation.

f. Quadrature. The basic reactance-tube circuit often is referred to as a *quadrature* circuit because the r-f voltage developed across its output terminals is leading or lagging the r-f current in its plate circuit by approximately 90°. When the injected reactance is capacitive, the current leads the r-f voltage; when it is inductive, it lags. If the plate resistance of the reactance tube is negligible in respect to the magnitude of the injected reactance, the phase angle approaches the value of 90°. However, since there is always a resistive component as well as a reactive one, this phase angle can never actually become the ideal quadrature relation of 90°.

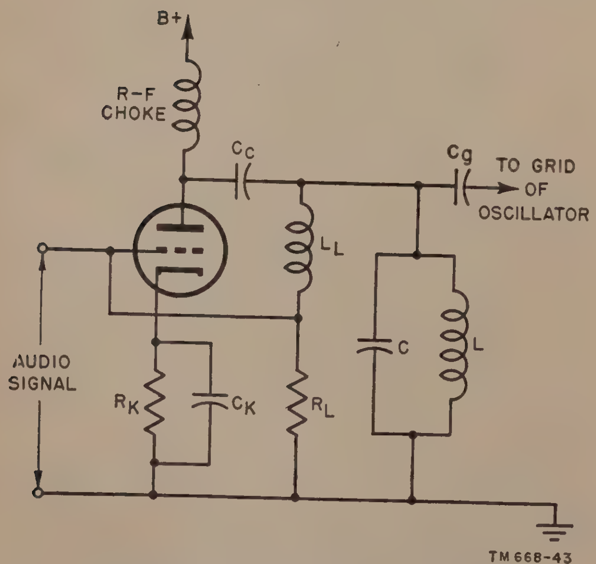


Figure 43. L-R reactance-tube load.

23. Practical Reactance Modulator

A of figure 45, shows a reactance modulator circuit used in a portable f-m transmitter, which injects a capacitance across the tank circuit of the master oscillator. For explanation purposes, the equivalent circuit of A is shown in B. The plate tank circuit of the master oscillator is represented by L_1-C_1 and the interelectrode capacitance between the grid and filament of the reactance modulator by C_{gf}. The interelectrode capacitance, C_{gf}, is connected effectively in series with $R75$ and in parallel with $L2$ to make up a phase-shifting network. Since

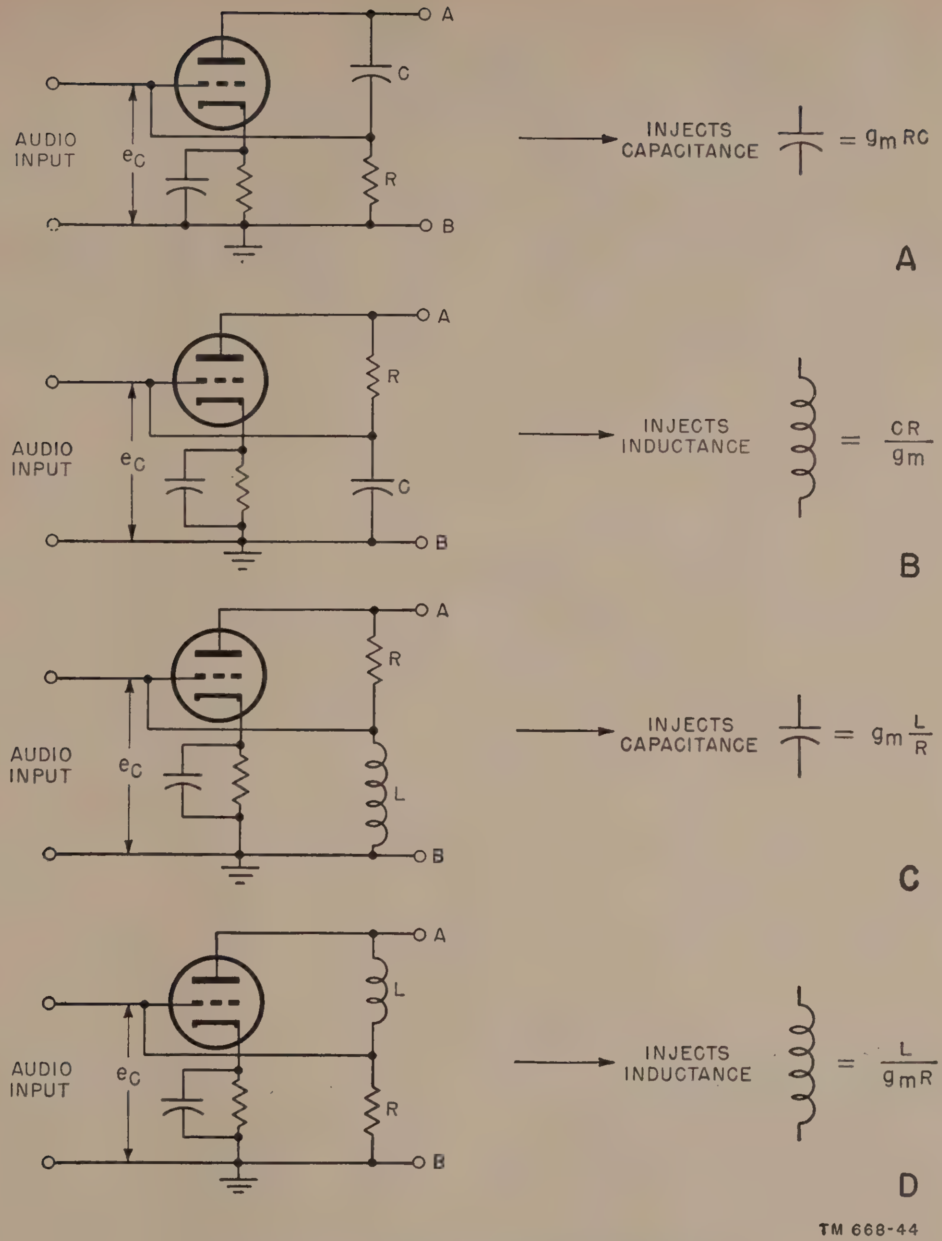


Figure 44. Four possible plate-load arrangements.

L_2 is coupled to L_1, the r-f voltage across it is 180° out-of-phase with the oscillator voltage across the tank. The resistance of $R75$ is made large in respect to the reactance of C_{gf}, and the current through C_{gf} is in phase with the induced voltage produced by $L2$. However, the voltage across C_{gf} lags the current through it by 90°, and this voltage is applied to the grid of the reactance tube. Since the grid voltage is in phase with the plate current, it lags the tank voltage and current by 180° plus 90°, or 270°. This is the same as saying that the plate-current leads the tank current by 90°. The effect is that of injecting a capacitance across $L1$-$C1$. Capacitor $C28$ blocks the d-c bias voltage on the oscillator grid from appearing on the reactance modulator grid. The purpose of $R17$ is similar to that of an r-f choke; it keeps the low-impedance microphone circuit from shorting the r-f components across C_{gf}.

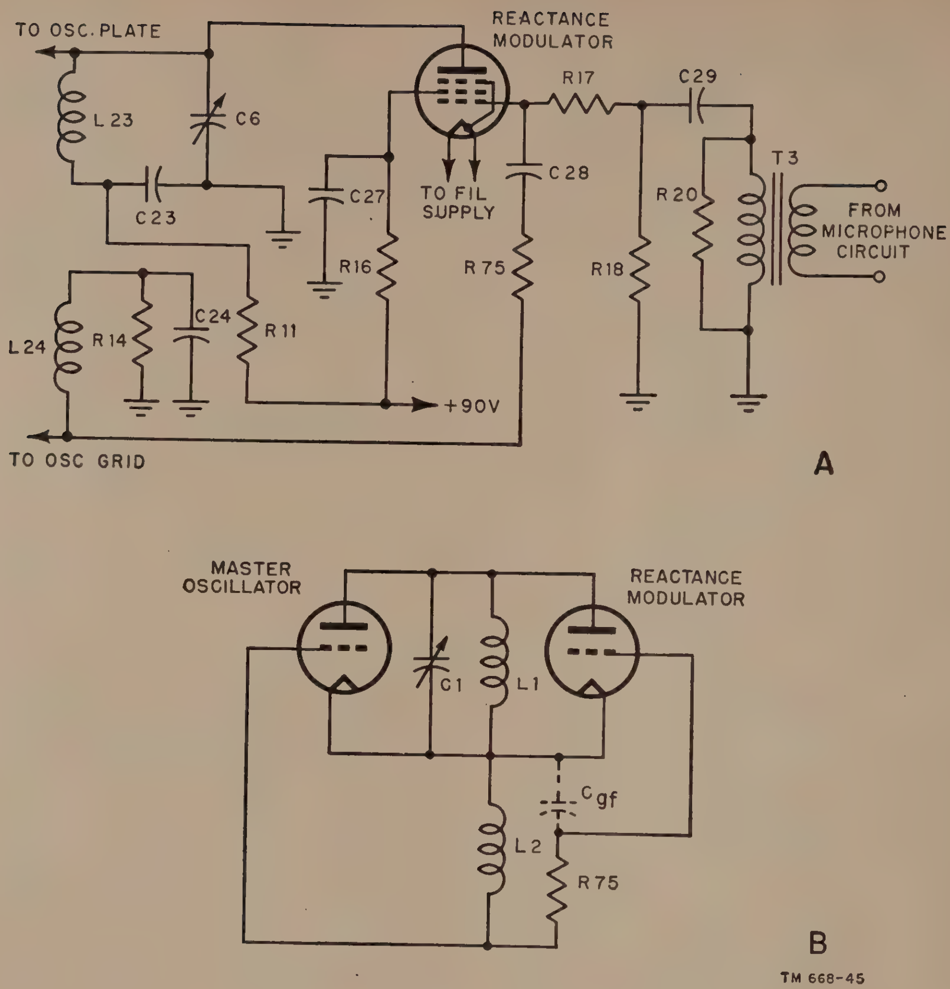


Figure 45. Practical reactance modulator.

24. Input Capacitance Modulator

a. In any vacuum-tube amplifier a definite impedance is present between the grid and the cathode. This impedance depends on the values of components in the external circuits of the other tube elements. Resistance in the plate circuit will result in the appearance of capacitance from grid to ground, depending on the amount of series plate resistance and the voltage gain of the tube. This is caused by the variations in electrical charge in the space between the grid and the plate. In the tube manual, a capacitance from grid to ground usually is listed for each tube under the heading of *input* capacitance. This value is determined with no resistance in the plate circuit. As the plate load is increased in value, the capacitance reflected into the grid circuit rises. How effective a given amount of plate load is in increasing this capacitance depends on the gain of the tube, and ultimately on the transconductance, $\mu = r_p g_m$. The changing of input capacitance with variations in plate-load resistance and tube transconductance is called *Miller effect.*

b. Since the input capacitance varies with the transconductance, it can change the frequency of an r-f oscillator when connected across the tank circuit. Therefore, Miller effect can be put to work in a direct frequency modulator (fig. 46). Operation is much like that of the reactance modulator, since variations in grid voltage at an audio rate are reproduced as changes in frequency of the oscillator. With no audio signal on the modulator, a fixed capacitance is injected across the oscillator tank. Consequently, the effective frequency of oscillation

goes down to the value set by the sum of the Miller-effect capacitance and the oscillator tank capacitance. When the audio signal is applied, the transconductance of the amplifier tube changes directly with the amplitude of that signal. The value of the Miller-effect capacitance increases and decreases with the amplitude variations of the audio signal. Since this capacitance is in parallel with the oscillator tank capacitor, the frequency varies above and below the audio-signal zero value, producing true f-m.

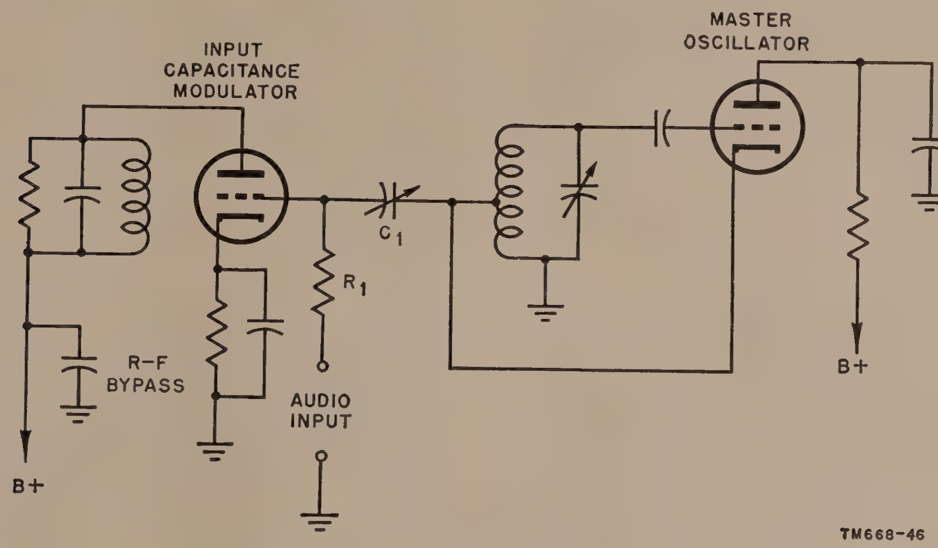


Figure 46. Input-capacitance modulator.

c. The amount of capacitance injected by Miller effect also depends on the internal tube capacitance between grid and plate. In a pentode, this internal capacitance is small, and the Miller effect is also small; in a triode, the internal tube capacitance itself usually is sufficient. To increase the amount of injected capacitance, a small capacitor can be connected externally between the grid and plate. A small value of plate load resistance is used to permit the tube to operate along a linear part of the plate current-grid voltage characteristic curve. Cathode bias is provided by a series resistor and bypass capacitor. To keep the deviation distortion low, the tube is coupled through $C1$ to a tap across a part of the oscillator tank. Audio voltage is applied to the grid of the tube through an isolating resistor, $R1$. The circuit is capable of a considerable amount of frequency deviation with low distortion. However, it is not as good for wide deviations as the reactance modulator. Furthermore, the amount of injected capacitance is not adjustable over the wide range that the quadrature circuit offers.

25. Diode Modulator

a. When a resistor and a capacitor are placed in series across a source of a-c voltage, the current flowing in the circuit is out of phase with the voltage. This current has two components, one caused by the resistance and one by the capacitive reactance. The resistive component is in phase with the applied voltage, and the reactive component leads the applied voltage by 90°. The resultant current therefore leads the applied voltage by some angle between 0° and 90°, the size of the angle depending on the relative size of the two components. If the resistance is large compared with the capacitive reactance, then the current leads the applied voltage by only a few degrees. If the resistance is small compared with the capacitive reactance, the current leads by nearly 90°.

b. If the resistance is made variable, the current can be made to lead the voltage by any value between 0° and 90°. If this variation can be controlled at the audio rate of a modulating signal, the resultant current can be used to frequency-modulate an oscillator. A diode is used as the variable resistance in the arrangement shown in figure 47, called a *diode modulator*.

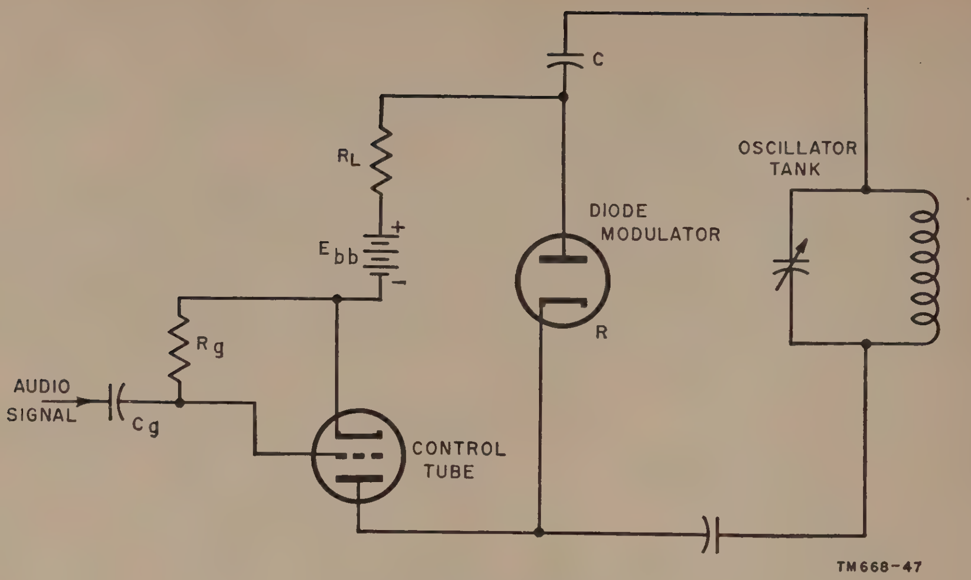


Figure 47. Diode modulator.

c. The diode and capacitor C are placed in series across the oscillator tank circuit. The reactance of C is made large in respect to the resistance of the diode. The applied voltage, E is the r-f voltage across the tank, and causes an r-f current to flow through the diode and C, which leads the voltage by nearly 90°. The plate current of the control tube also flows through the diode and R_L to the power supply. With zero-modulating signal this current is constant, and the modulator injects a fixed reactive current into the tank which determines the operating frequency.

d. When an audio-modulating signal is applied to the grid through C_g, an audio current flows through the plate circuit of the control tube. This current increases and decreases in step with the audio voltage. Since the current through the control tube also flows through the diode, the current flow through the diode also increases and decreases. This has the same effect as increasing and decreasing the resistance of the diode with a constant voltage applied. The *effective* resistance of the diode in series with the r-f current, therefore, is varied at the audio rate. The variations in the resultant reactive current injected into the tank circuit change the frequency of oscillation. A change in the reactive current has the same effect as a change in the capacitance across the tank circuit, and the result is a frequency-modulated signal.

e. Less noise and distortion are present in the diode modulator than in the conventional reactance-tube modulator. In addition, it is easy to set the operating range so that the frequency deviation is more nearly proportional to the modulating signal. The reactance-tube modulator has considerable variation in the resistive current drawn during modulation which always introduces some amplitude modulation. In the diode modulator, the actual resistive change that takes place across the oscillator tank is so small that there is very little resultant distortion. The reactance modulator, however, is capable of a much greater maximum frequency deviation.

26. Frequency Modulation of R-C Oscillator

a. Several communication systems have been discussed in which the information transmitted was not necessarily speech. One of the most common of the nonspeech applications of f-m is facsimile, involving the transmission of pictures, maps, and similar material which are printed automatically at the receiver in response to signals from the transmitter. Facsimile often uses a system of f-m known as *sub-*

carrier modulation which consists of the amplitude modulation of an r-f carrier by an audio tone. The a-m wave then is frequency-modulated and serves as a carrier for the applied signal.

b. Subcarrier modulation requires wide deviation relative to the carrier frequency. For example, if the subcarrier is 4,000 cps, it is common to find deviation which varies the frequency between the limits of 2,000 and 6,000 cps. This variation is more than 50 percent of the subcarrier frequency itself. Such relatively wide deviation cannot be obtained from a reactance modulator driving an ordinary *L-C* oscillator without excessive distortion. The variation of transconductance necessary to produce the relatively enormous change in injected capacitance or inductance needed cannot be obtained from the straight-line part of the e_c-i_b characteristic. Therefore, a completely different solution has been worked out, using an oscillator that has no inductance in its circuit.

c. Oscillators for the audio-frequency range can be made using only resistors and capacitors. Their operation is based on the principle that a series combination of a resistor and a capacitor causes a definite phase shift. When the values of the resistance and the capacitive reactances coincide, the current in the circuit leads the applied voltage by 45°. If the voltage developed by the current in the resistor is applied to a second identical series network, the current in this circuit leads the original input circuit voltage by 90°. If the current is made to flow through two more identical networks (fig. 48), the overall phase shift from the input to the output of the network (4 resistors and 4 capacitors) is 180° at a particular frequency.

d. The plate voltage of a vacuum tube is opposite in polarity to that of the voltage on the grid, and these voltages can be considered 180° out of phase with each other, when the plate load is resistive. A decrease in grid voltage causes a corresponding decrease in plate current. If this decreased plate current flows through a load resistance, the voltage drop across that resistor decreases, and the plate voltage rises toward the supply voltage. If part of the plate voltage is returned to the grid through the network, shown in A, the applied grid signal is in phase

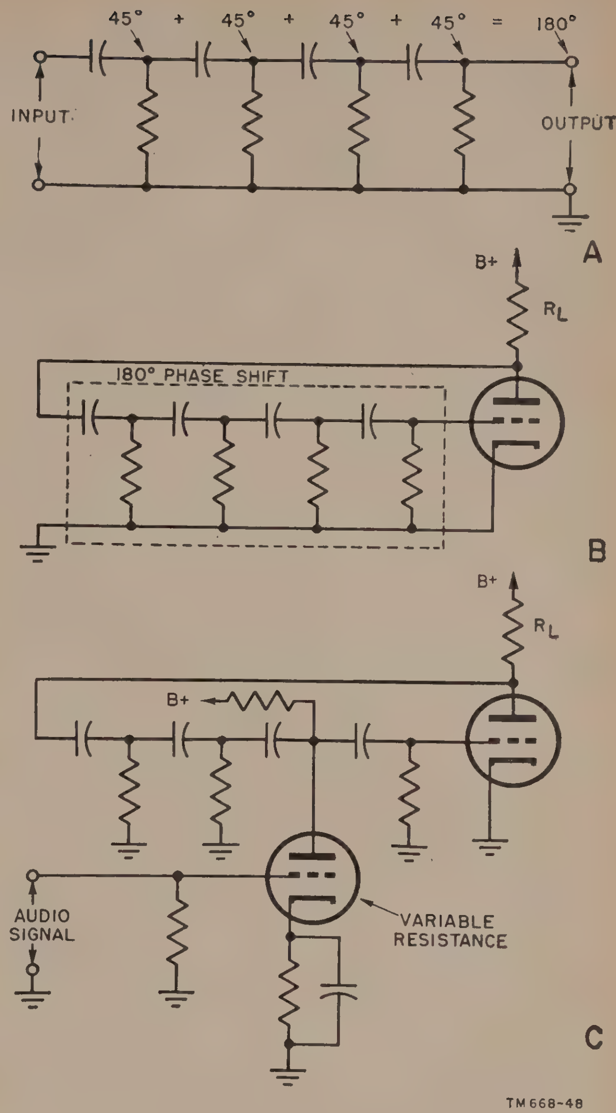


Figure 48. Development of f-m phase-shift oscillator.

with the output of the network. The result is an increase in signal input to the network with correspondingly greater increase in plate voltage, until the accumulative action causes the circuit to break into oscillation. This is the arrangement in the *phase-shift oscillator* in B.

e. The phase-shift oscillator is capable of generating a wide range of frequencies throughout the entire audio range. When the capacitive reactance is such that the phase shift in each section is exactly 45°, the greatest tendency for oscillation exists. Any slight variation in plate current or supply voltage starts the circuit os-

cillating. The frequency of oscillation can be changed by altering the phase shift of any or all of the components of the network. If the resistors are made variable, the frequency can be changed over a considerable range. The amplitude of oscillation is comparatively independent of the losses in the network, provided that the gain of the tube exceeds a certain critical value. Changing the value of one or more of the resistors alters the loss in the network, but does not alter the amplitude appreciably, although it does change the frequency. Therefore, wide frequency excursions can be obtained through changing the resistor values with no variation in amplitude of oscillation. This means that a frequency-modulated signal can be generated which has very little distortion.

f. Instead of varying the resistors mechanically, an electronic circuit has been devised. This is shown in C of figure 48. One of the resistors of the phase-shift network is replaced by the equivalent plate resistance to ground of a vacuum tube. This is accomplished by connecting the plate to the phase-shift network and the cathode to ground. Plate voltage for the tube is supplied through a suitable load resistor. When the grid voltage of the tube is varied by modulating signal, the plate resistance changes over a wide range. In fact, when the tube goes to cut-off, the resistance is extremely high. Similarly, when the grid voltage is zero, the plate resistance is very low. The resistance of this element of the phase-shift network, consequently, can vary over a range which is directly proportional to the modulating signal. This variation of resistance changes the over-all phase shift of the network so that feedback is developed at a different frequency. The variation of plate resistance of the modulator tube is directly proportional to the amplitude of the modulating signal, and the frequency of the phase-shift oscillator is also proportional to the control resistance over a wide range. It follows that the frequency of the oscillator also must be proportional to the amplitude of the modulating signal over an equally wide range. This is the characteristic desired for producing f-m in the subcarrier modulation process.

27. Permeability-Modulated Direct F-M

a. An extremely simple frequency-modulator circuit has been devised for use in very compact portable equipment, where saving an extra tube means an increase in the service life of the batteries. This modulator needs no additional tubes for production of f-m. The operating principle is based on the variation of inductance in an iron core coil with changes in the magnetic flux passing through it. Inductance can be altered by changing the permeability of the iron core—that is, the ease with which the magnetic field passes through it.

b. If the iron core is made a part of an audio transformer in the plate circuit of an audio amplifier, variations in the plate current alter the permeability of the core. This, in turn, changes the inductance of a part of the coil used to form the tank inductance of an r-f oscillator. A fair amount of deviation can be produced in this way without any extra modulating tubes. The system has the disadvantage of high distortion and low deviation capabilities, since the powdered-iron core must be operated in a highly saturated magnetic condition.

Section II. INDIRECT METHODS OF PRODUCING FREQUENCY MODULATION

28. Phase Variation

a. General.

(1) All of the frequency-modulation circuits discussed in section I consisted of oscillators using reactive elements in their frequency-determining networks. Such circuits, however, are inherently unstable in respect to frequency. For very critical applications where it is desired that stations operate on precisely specified frequencies, crystal oscillators are used. A quartz crystal is the electrical equivalent of a resonant circuit with an extremely high *Q*. The resonant frequency is determined by the mechanical properties of the quartz itself.

(2) The reactance changes that can be produced with a direct-method frequency modulator have little effect on the resonant frequency of a crystal oscillator.

Therefore, this oscillator is considered to be fixed in frequency. If a crystal oscillator can be phase-modulated, the output can be converted easily to frequency modulation. Indirect f-m is produced by using a suitable correcting network to convert a phase-modulated signal.

b. Phase Variation. A change in the phase of a signal can be produced by passing the signal through a network containing resistance and reactance. When a voltage is applied to a capacitor and a resistor in series, for example, the current leads the applied voltage by an amount dependent on the relative values of resistance and capacitive reactance. This current develops a voltage across the resistor which leads the applied voltage. If the series combination is considered to be the input, and the output voltage is taken from across the resistor, a definite amount of phase shift is introduced. If the fixed-frequency signal from a crystal oscillator is passed through this network, its phase at the output is shifted by an amount depending on the ratio of the reactance to the resistance. If the resistor can be varied, the phase angle of the network changes to correspond with the newly established ratio of reactance to resistance. When the resistance is varied with an applied audio signal, the phase angle of the output changes in direct proportion to the audio-signal amplitude and produces a phase-modulated signal.

c. P-M to F-M. The basic circuit of a phase modulator with the resistor replaced by the variable plate resistance of a vacuum tube is shown in figure 49. The plate resistance of the triode changes with grid voltage and therefore serves as the variable resistor. Since the plate resistance of the triode varies with the audio signal applied to the grid circuit, the phase between the input of the circuit and the output changes with the audio signal. As the grid swings positive, the plate resistance drops and the phase angle of the output increases; when the grid swings negative, the plate resistance rises and the phase angle decreases. The change of plate resistance with various values of grid voltage is exactly proportional to grid voltage over a small range. If the phase angle of the network is changed between *wide* limits, the *amplitude* of the output changes. This means that the modulator can produce only a limited phase deviation without distortion. In general, it is only reasonably good over a range of less than 25° of phase shift.

29. Constant-Impedance Phase-Shift Modulator

Since the resistance of the modulator tube varies, the voltage across it varies in amplitude

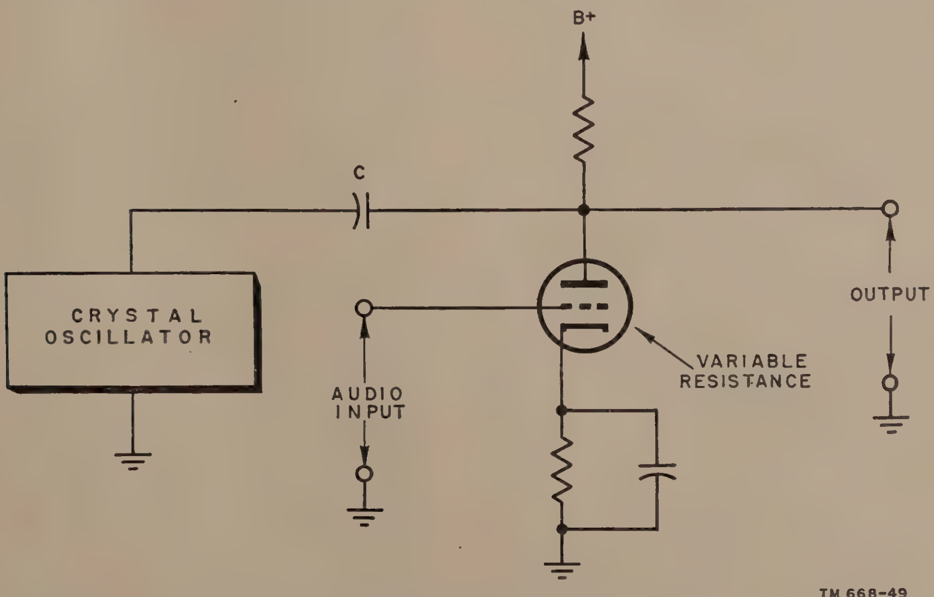


Figure 49. Simple vacuum-tube phase modulator.

as well as in phase. The impedance of the entire phase-shift network varies and introduces distortion in the amplitude of the output signal. This undesirable effect can be overcome, however, if a constant-impedance network is used. A phase modulator having a constant-impedance network is shown in figure 50. The impedance across the cathode resistor, R_k, is used since it changes with variable grid voltage and is more uniform than changes in plate resistance. The resistance between cathode and ground changes with the audio signal, and the phase of the output signal is modulated as desired. The inductor, L, serves to correct any change in the total impedance, keeping the amplitude of the output constant. For any change in frequency, a change in capacitive reactance is canceled by an opposite change in inductive reactance.

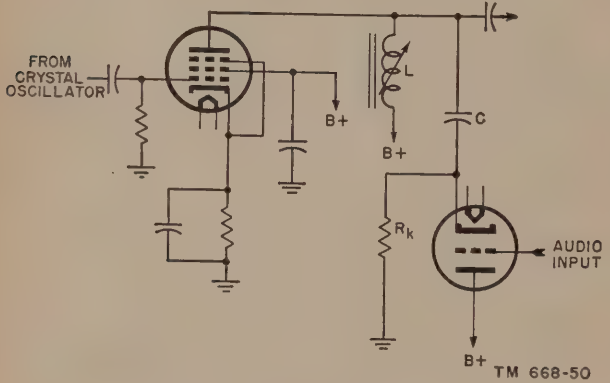


Figure 50. Basic circuit of constant-impedance phase-shift modulator.

30. Audio Correction

a. The equivalent frequency deviation in p-m is proportional to the audio-modulating frequency. This effect is undesirable in indirect f-m production, since the frequency deviation must depend *only* on the amplitude of the audio signal. To convert into f-m the phase-modulated signals produced by the circuits previously described, it is necessary to pass the audio through a correction network to shift its phase 90°. A simple *R-C* series network, with the output voltage taken across the capacitor, can accomplish the desired results. If the resistance is much larger than the reactance of the capacitor at the lowest audio frequency, the current flow is determined primarily by the resistance. Therefore, the current is relatively constant, since the resistance is constant. The reactance of the capacitor changes with frequency, but its effect on the total current flow is small because the reactance decreases as frequency increases and current flow is limited by the high resistance. The voltage across the capacitor, therefore, is equal to the relatively constant current, multiplied by the changing reactance.

b. Since the capacitive reactance is inversely proportional to the frequency and the resistance is fixed, the output voltage is proportional to the reactance alone. Therefore, it also is inversely proportional to frequency, as desired. This type of circuit sometimes is called a *pre-distorter* or *audio-correction* network because it changes the response of the phase modulator so that indirect f-m is produced instead of direct p-m. Since signal amplitude is lost in this type of network, the output sometimes is amplified before it is applied to the phase modulator.

31. Link Phase Modulator

a. The variations in transconductance of a tube with a varying audio signal can be made the basis for a phase modulator. In A of figure 51, the oscillator voltage is applied to the grid circuit of the modulator through coupling capacitor *C*8. The oscillator signal can reach the plate by two paths. One is through the grid-plate capacitance of the tube shown in the diagram as C_{gp}. The other is provided by the transconductance and represents the normal operation of the stage as an amplifier. The plate voltage of an amplifier is always 180° out of phase with the grid voltage. However, the voltage fed to the plate through the grid-plate capacitance is always in phase with the grid voltage. It is therefore 180° out of phase with the amplified voltage.

b. If the tube is operated as a normal high-gain voltage amplifier, the amplified plate voltage is much larger than that caused by the grid-plate capacitance. When a large unbypassed cathode resistor is used, the operating point of the tube can be displaced so that the transconductance and amplification are greatly reduced. The omission of the cathode bypass capacitor permits the variations in plate current to establish degeneration—that is, a varying voltage on the cathode that acts in opposition to effective

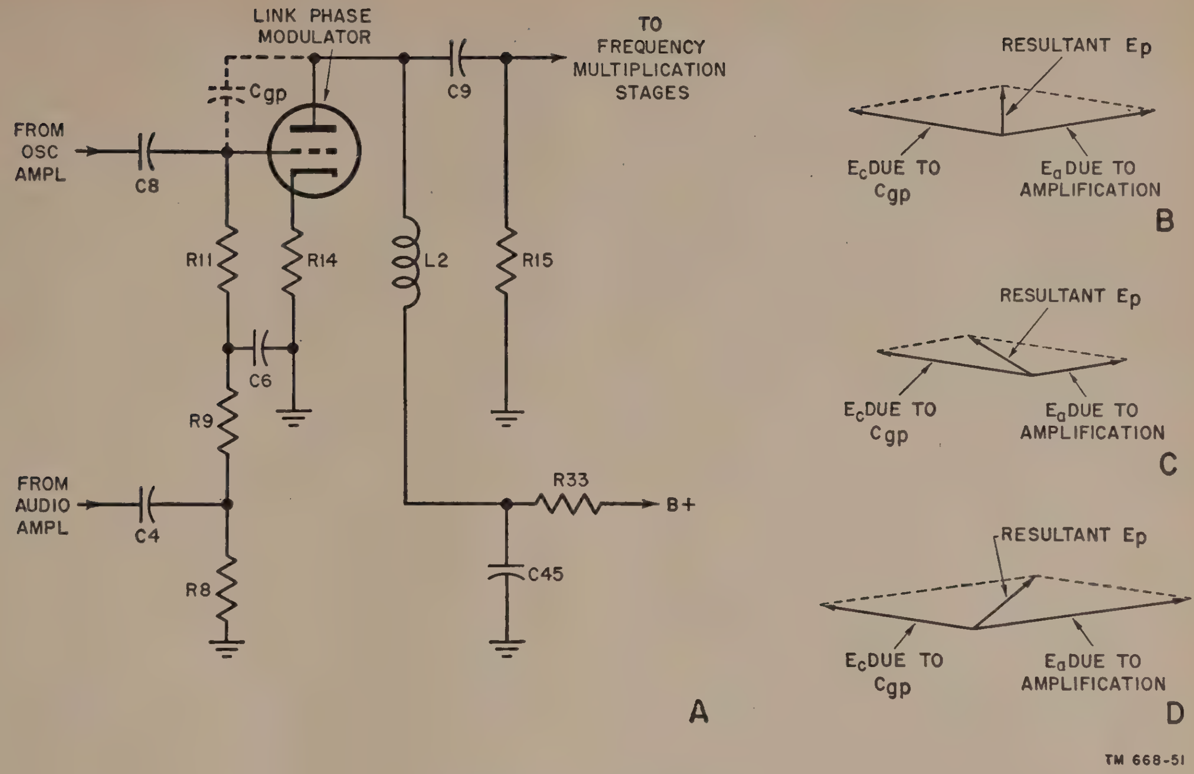


Figure 51. Phase modulator.

grid input voltage—further reducing the transconductance. Because of the grid-plate capacitance, the amplified voltage component on the plate can be reduced in this way to approximately the same magnitude as the component.

c. Resistor *R*8 and the series combination of *R*9 and *R*11 form the grid circuit of the modulator. The plate voltage is applied to the tube through the plate load inductor *L*2, which is broadly resonant with stray capacitances at the operating frequency. The stage is coupled capacitively to the following stage by capacitor *C*9. Capacitor *C*6, in conjunction with the resistance of *R*9, forms the audio correction network.

d. The vector diagrams in B illustrate the relative values and angular relationships of the amplified and capacitive voltage components, E_a and E_c, at the plate. The instantaneous a-c plate voltage is the vector sum of E_a and E_c as shown by the resultant vector, E_p. Audio voltage is applied to the grid of the tube through resistors *R*9 and *R*11, which serve to isolate the r-f and audio circuits. The transconductance is varied at an audio rate by the modulating signal. Consequently, the amplified component of plate voltage varies in amplitude. As the audio signal becomes positive, this plate voltage component is decreased in amplitude but the signal coupled through the grid-plate capacitance from the oscillator does not change in either amplitude or phase. Its vector, E_c, remains constant, as shown in C. Since the amplified voltage, E_a, changes and the capacitive voltage, E_c, does not, the amplitude and phase angle of the resultant, E_p, must change. When the grid of the tube is made negative by the applied audio voltage, the amplified plate voltage increases. The resultant voltage, E_p, changes in amplitude and phase angle, in a direction opposite to its change when the grid swings positive. This is shown in D. The phase of the output signal, therefore, varies in phase, in accordance with the amplitude of the input signal.

e. There are amplitude changes in addition to the phase changes that take place in the r-f voltage on the plate. These are not very great, however, and they can be eliminated in the

stages following the modulator stage. Very little amplitude variation appears in the output of the transmitter.

f. This system is not capable of a phase shift any greater than the approximate 180° separating the amplified and capacitive voltages. In fact, the curvature of tube characteristics at low levels of amplification does not permit the variation in amplified plate voltage to be exactly proportional to the applied audio voltage over its entire range. The actual phase deviation for reasonably low distortion must be much less than 180°. In general, peak phase deviations for the lowest audio frequency to be passed (generally 300 or 350 cps) are about 75°.

32. Sonar Phase Modulator

a. Some of the disadvantages of the Link circuit are overcome in the circuit of figure 52. The oscillator voltage appears on the plate through the dual paths of the tube and the grid-plate capacitance, as before. The resultant instantaneous a-c plate voltage with no modulation is approximately 90° out of phase with the applied r-f grid voltage. When the transconductance is varied by the audio signal, the amplified voltage component varies, and therefore the phase of the resultant also changes.

b. The plate load coil, however, is divided into two parts and the plate-supply voltage is introduced at a tap. The two parts of the coil are 180° out of phase with each other. Part of the out-of-phase voltage from the plate is fed back to the grid through capacitor *C*. The coil is tapped so that the turns ratio between the upper and lower ends is approximately 2 to 1. This provides an adjustment for the phase and amplitude of the voltage acting between grid and plate, by varying the feedback through capacitor *C*. Capacitor *C*1, across the upper part of the coil, varies the magnitude and phase angle of the impedance in the plate circuit through resonance. These circuit modifications appreciably increase the available phase shift by keeping distortion low.

c. When the modulating signal voltage is applied to the grid, there is variation in the instantaneous a-c plate current of the tube. Because the coil has a powdered-iron core, the variations in plate current change the magnetic saturation (par. 27). This changes the actual inductance of the coil. With the variation of this inductance caused by the audio signal, the resonant frequency and the phase angle of the plate load circuit also are varied at an audio rate. This phase variation adds to that already produced by the fundamental circuit, and to the variation produced through adjustment of the feedback circuit. The resultant frequency deviation is approximately six times that of the Link phase modulator.

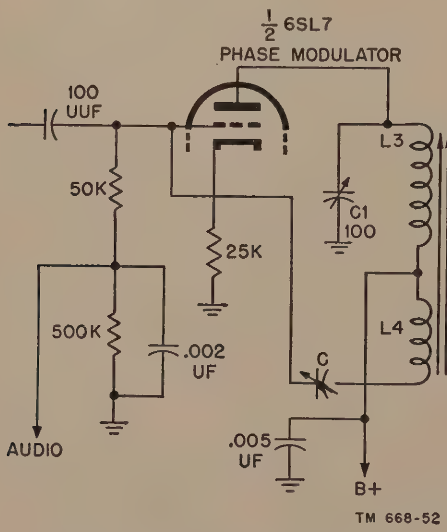
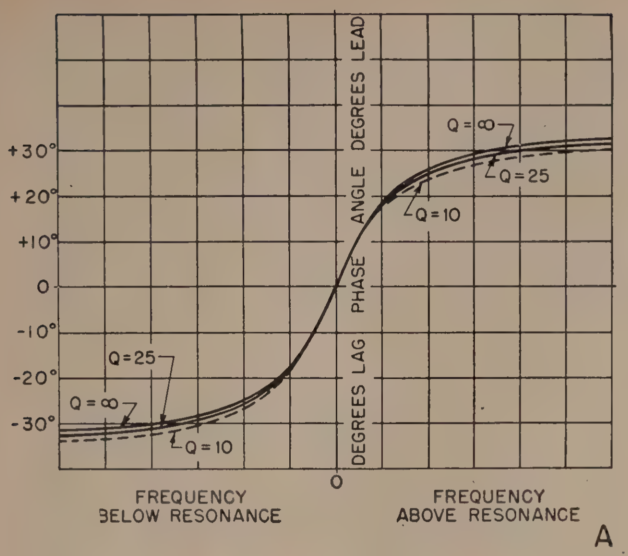


Figure 52. Sonar phase modulator.

33. Reactance-Tube Phase Modulator

a. It is possible to produce phase modulation by connecting a variable reactance across the resonant load circuit of an r-f amplifier (fig. 53). The variation in the phase angle of a parallel resonant circuit shown in A, is plotted in respect to the frequency. As the frequency of the voltage applied increases, the capacitance in the circuit begins to be predominant, and the resultant total current leads the applied voltage. When the frequency of the applied voltage is lower than the natural resonant frequency of the tuned circuit, the inductance has a lower reactance than the capacitor and, consequently, draws the major share of the current. Therefore, the total current lags the applied voltage.

b. If the frequency of the resonant circuit is varied and the applied voltage kept constant, the same variation of phase angle is produced.



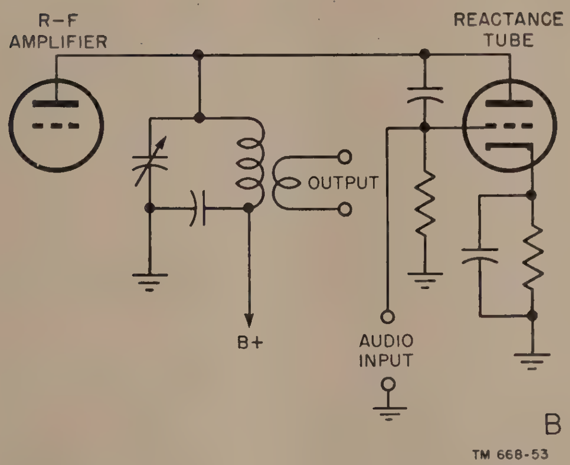


Figure 53. Reactance-tube phase modulation.

The curve in A also applies, except that the center frequency is that of the applied voltage, whereas the frequencies on either side are those to which the resonant circuit is tuned. The circuit can be tuned by changing the value of either the inductance or the capacitance. Changing either at an audio rate with a reactance modulator produces phase modulation. The shape of the phase variation curve depends on the *Q* of the tuned circuit. This, in turn, depends almost entirely on the construction of the inductor. Therefore, injection of capacitance usually is used to avoid changes in the shape of the curve.

c. The reactance modulator, in B, is designed to inject a variable capacitance across the resonant-circuit load of an r-f amplifier. Note that plate voltage for the modulator and the amplifier are supplied in common. The injected capacitance changes the tuning of the tank circuit in response to the variations in audio signal. This, in turn, changes the phase angle of the current drawn by the tank. Consequently, the phase angle of the output from the tank circuit also varies. The curve in A shows that the change in phase is proportional to the detuning over a small range approximately between the limits of —25° and +25°. This circuit, therefore, cannot produce much phase deviation. It has the advantage of permitting the use of a reactance tube with a crystal oscillator-amplifier combination. However, the wide deviations associated with the reactance modulator when used with a self-excited oscillator cannot be obtained.

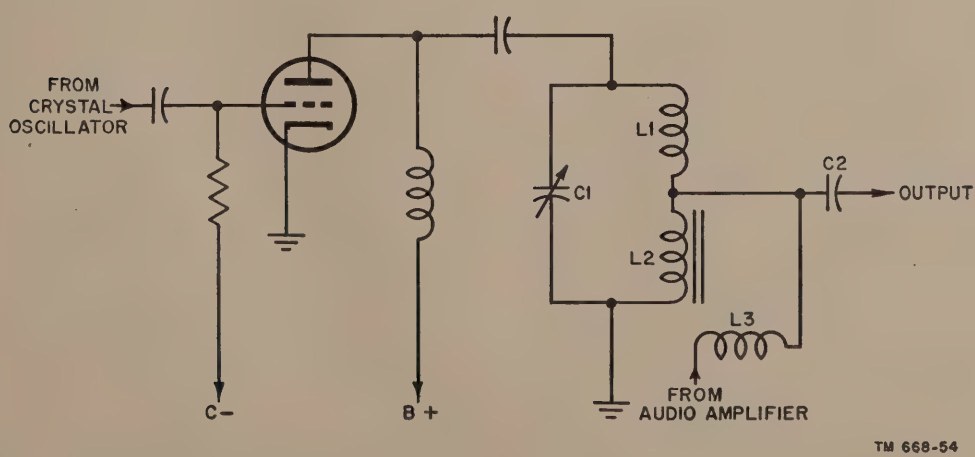


Figure 54. Nonlinear coil modulator.

34. Nonlinear Coil Modulator

a. It is possible to construct a coil which will have the property of introducing phase modulation into a carrier, when both radio and audio frequencies are passed through it. This is termed a *nonlinear coil modulator* (fig. 54). The output from the r-f amplifier is passed through a plate load consisting of the resonant circuit *C*1, *L*1, and the special nonlinear coil, *L*2. The output of the audio amplifier is applied through *L*2 through *L*3, and the phase-modulated signal produced is coupled to the output by *C*2. The rest of the circuit is a conventional r-f amplifier. The nonlinear coil circuit produces a frequency deviation of nearly 1 kc. This phase-modulator circuit is relatively efficient in terms of the amount of initial phase deviation.

b. Normally, when a current is passed through a coil without a magnetic core the current that flows is of the same waveshape as the voltage applied. If, however, a magnetic core is introduced into the coil, this situation changes. When a magnetic field exists in a magnetic material, there is a definite magnetizing force corresponding to that field. As the field increases in strength, the material becomes magnetized until a point is reached where the increase in the magnetizing force produces no increase in the magnetic field set up. When the material is fully magnetized and the magnetic flux cannot increase, a state called *saturation* is reached.

c. When current flows in a coil, it sets up a magnetic field that magnetizes the core material placed in the immediate vicinity. As the current increases, the corresponding magnetic field increases and, also, the magnetization of the core. Because of saturation, however, there is a definite point beyond which any additional current causes no additional magnetization. Special alloys, such as permalloy, when used as cores reach the saturation point at very low values of the magnetizing field. A coil wound around a permalloy core reaches saturation with a very small amount of applied current. When a sine-wave voltage is applied to such a coil, the magnetic flux increases rapidly until the core saturates, and the flux then becomes relatively constant.

d. The relation between the applied magnetizing current and the voltage developed across the coil is shown in figure 55. When the current begins to increase toward its maximum value, the magnetization of the core rises rapidly, with a rapid increase in flux. While the current flow is above saturation (A to B), there is no change in the flux, since the core is saturated. The same situation occurs on the negative half-cycle between points C and D.

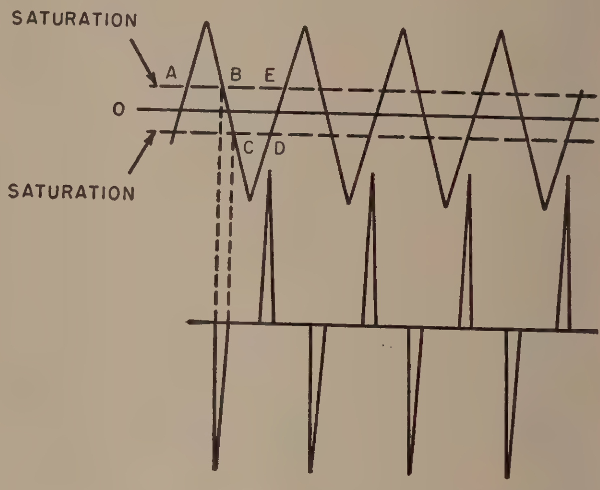


Figure 55. Voltage pulses developed across nonlinear coil.

e. The induced voltage depends on the rate of change of the flux. When the flux is not changing, no voltage is induced. In a coil wound on a permalloy core, the flux is changing rapidly during part of the cycle and, during that time, large voltages are induced in it. These voltage pulses of high amplitudes occur *only* during the periods of rapidly changing flux. At other times, when the flux is nearly constant, little or no voltage is induced across the coil. It is shown in the figure that these voltage changes take place exactly 90° after the current peaks. The polarity of the pulses depends on the direction of the magnetic flux. Therefore, on opposite half-cycles of magnetizing current, the pulses are of opposite polarity. This 90° difference is constant in respect to the magnetizing current, and, since this current is supplied by an r-f oscillator, the pulse is constant in frequency.

f. Assume, however, that, in addition to the r-f energy, audio signals are simultaneously

applied to the nonlinear coil. They have the same magnetizing effect on the core material and they combine with the r-f current to produce the current wave, as shown in the first three lines of figure 56. Curve A represents the current in the coil, caused by the carrier r-f. Curve B is the current produced by the modulating signal, assumed to be sinusoidal, for simplicity of analysis. The combined waveform is shown in line C. It is clear that the resultant current no longer goes through the zero axis in the same time interval as before, and, therefore, the region of maximum rate of change of flux is different for each cycle and depends on the audio voltage. These combined currents produce voltage pulses across the coil at different instants during the audio cycle, as in D. The variation in the level of the r-f current at different points of the r-f cycle causes this effect. Sometimes, the pulse is produced at the normal interval of the unmodulated carrier. At other times, the pulses are spaced more, or less, than 360° part. These variations in the spacing of the pulses, with different values of modulation voltages, are obviously equivalent to displacements in the relative phase of the pulses. In other words, a change in the a-f voltage shifts the phases of the pulses in respect to the phase of those pulses produced by the unmodulated carrier. Therefore, these pulses are effectively phase-modulated.

g. If the phase-modulated pulses of voltage derived from the nonlinear coil are applied to a rectifier and limiting amplifier, only the pulses of one polarity will be passed. Furthermore, the limiting action will reduce the slight variations in the amplitude of the pulses that appear in D. The action of this rectifier and limiter is shown in E. When pulses of sharp amplitude pass through a resonant circuit, they set it into oscillation at its natural resonant frequency. If these pulses from the output of the rectifier pass through a resonant circuit, which is tuned to their repetition frequency, a sine wave is produced. Since the pulses are phase-modulated by the audio voltage, the resultant sine wave also will be phase-modulated.

35. Balanced Modulator

a. Vector Relations Between Amplitude and Phase Modulation—Amplitude Modulation. In the earlier chapters, it was pointed out that when the modulation index of an f-m signal is less than .5, only one effective side-band pair is produced. Therefore, the side bands of an f-m wave with a low modulation index are similar to those of an a-m wave. The major difference between the side bands of a normal a-m wave and those of an f-m signal with a small deviation lies in their phase relationship to the carrier. An a-m signal consists of a carrier and two side bands for each audio frequency present in the modulating wave. For an f-m wave with modulation index of less than .5 the same holds true. If the two side bands of the a-m signal are added together vectorially, a new wave results, called a *double side-band wave.* It represents the difference between the frequency of the modulated wave and that of the carrier. For a narrow-band f-m signal, a similar double side-band wave also is formed when the two side bands are added. The a-m and f-m double side-band waves are ordinarily identical with one another, if they are of the same frequency and amplitude.

b. Double Side-Band Vectors. In the vector diagram shown in A of figure 57, the vector, *AB*, represents the amplitude of the unmodulated carrier. Vectors *BC* and *BE* are the two side-band components. *BD* is the vector sum of the side bands, or the double side-band vector. The resultant a-m wave therefore is the vector, *AD*. Each of the side-band vectors rotates about the tip of the carrier vector with a rotational frequency equal to the difference in frequency between the carrier and each side band, respectively. Consequently, the two side-band waves, which are equally spaced in frequency on either side of the carrier, rotate at the same speed but in opposite directions. One is higher in frequency, producing a positive difference, whereas the other is lower than the carrier frequency, producing a negative difference. The resultant of the two side bands, therefore, must move on the same line as the vector representing the carrier frequency, if it starts in phase with the carrier. This either adds to or subtracts from the amplitude of the carrier, as shown in B and C. In B, the two vectors add to produce a resultant that adds to the carrier amplitude. In C, they add in the opposite direction, and therefore subtract from the carrier vector. This

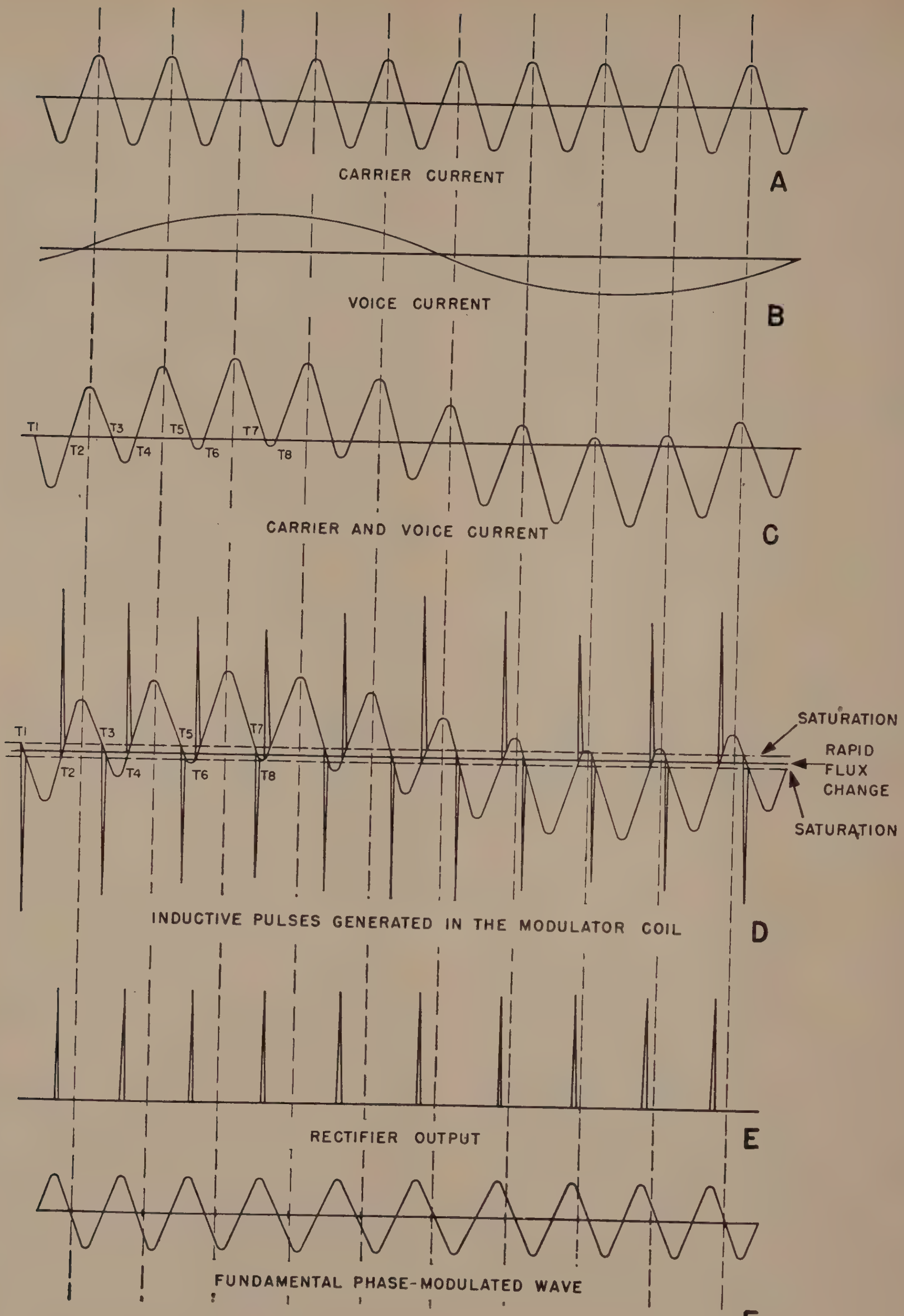


Figure 56. Waveforms developed in nonlinear-coil modulator.

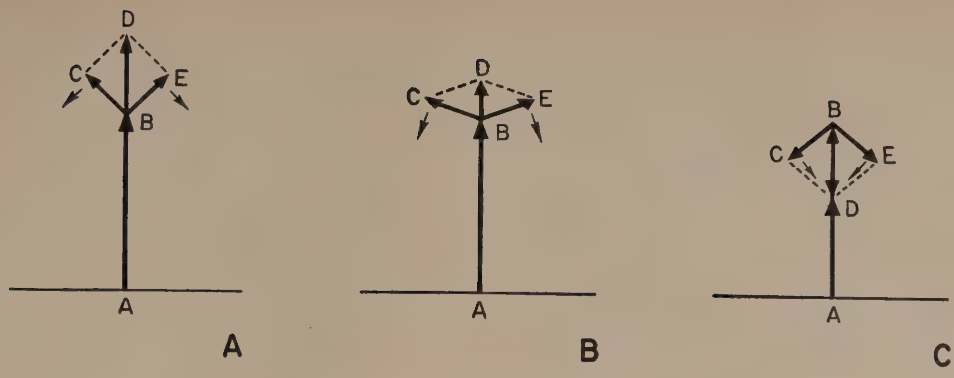


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Figure 57. Vector relations in double side-band a-m system.

causes changes in the amplitude of the combined signal, but there are no phase changes in the carrier itself. The resultant of the combined side-band signals is in phase with the carrier frequency.

c. Vector Relations Between Amplitude and Phase Modulation—Phase Modulation. In the case of phase modulation, when the frequencies and amplitudes of the side bands are the same as those in a-m (modulation index less than .5), the vector relations of the side bands and the carrier are different. Specifically, the double side band, which results from the addition of the two rotating side-band vectors, is such that it is always 90° out of phase with the carrier, as in A of figure 58. Instead of starting in phase with the carrier, as they do in a-m, the side-band vectors begin to rotate in opposite directions, but 90° out of phase with it. Consequently, the double side-band vector is always 90° out of phase with the carrier. It can be either on the right side of the carrier, as in B, or on the left side, as in C. The resultant of the sum of the two side bands and the carrier (vector *AD*) therefore shifts alternately from one side of the unmodulated carrier position to the other. It is undergoing a variation in respect to the unmodulated phase condition of the carrier, and, consequently, p-m is produced. When the resultant, *AD*, is on the right of the carrier, an angle, $\phi 1$, is generated. As the side bands rotate, and *BD* becomes smaller, vector *AD* changes, generating angle $\phi 2$. Similarly, $\phi 3$ is the generated angle, when the double side-band resultant is on the lower side of the carrier.

d. Changing A-M to P-M. When the deviation is low, a-m, f-m and p-m signals are all similar, with the exception of the phase relationship of the double side band to the carrier. This indicates a possible way to generate an indirect f-m signal from an a-m wave. A signal is amplitude-modulated with a low percentage

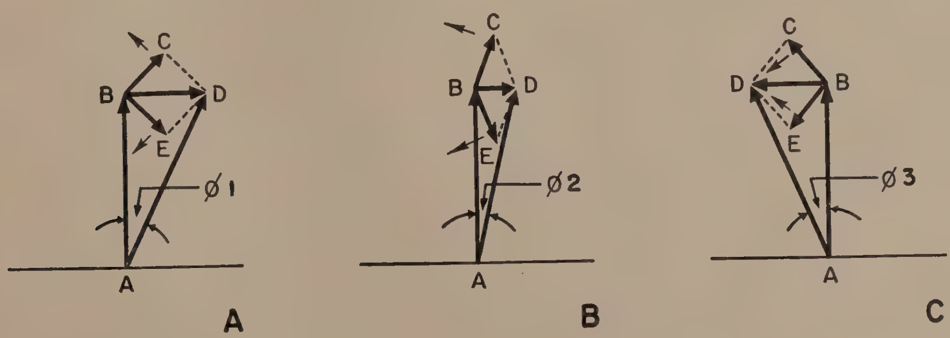

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Figure 58. Vector relations in a double side-band f-m system.

of modulation. Then it is passed through a circuit which removes the carrier-frequency component, leaving only the double side band. This double side band is shifted in phase by 90°. If it now is recombined with the carrier component, the result is a phase-modulated wave. This amounts to converting the vector diagram in A of figure 57 into that of A of figure 58 by rotating the double side band 90°. When this is done, the resultant vector varies in phase and in amplitude simultaneously, in respect to the carrier frequency. The slight amplitude variation remaining can be removed in a limiting amplifier, as explained previously.

e. Indirect F-M System: A-M to P-M.

(1) A system for the production of f-m by the indirect method is shown in figure 59. A crystal-controlled oscillator generates a low-frequency carrier that is fed into an r-f amplifier. At the same time, it passes through a device called a balanced modulator which amplitude-modulates it with an audio wave that has been passed through an audio correction network, producing the two side bands and simultaneously removing the carrier itself. The output from the balanced modulator then is fed through a network that shifts the phase of the double side band by 90°.

(2) The amount of amplitude modulation developed by the balanced modulator is small and the resultant side bands are of low amplitude. The double side band, after passing through the 90° phase-shift network, is recombined with the carrier frequency as derived directly from the crystal oscillator. This results in a signal that is almost pure p-m, because the original amplitude variations of the double side bands are small in comparison with the unmodulated carrier. The amount of phase deviation possible with this system is small because of the low amplitude of the double side bands. However, when phase deviation is kept small, distortion in the modulation is extremely low.

f. Balanced Modulator.

(1) The heart of the f-m system outlined above is the circuit that modulates the carrier wave, producing the double side band while eliminating the carrier itself. There are several forms of such balanced modulator circuits,

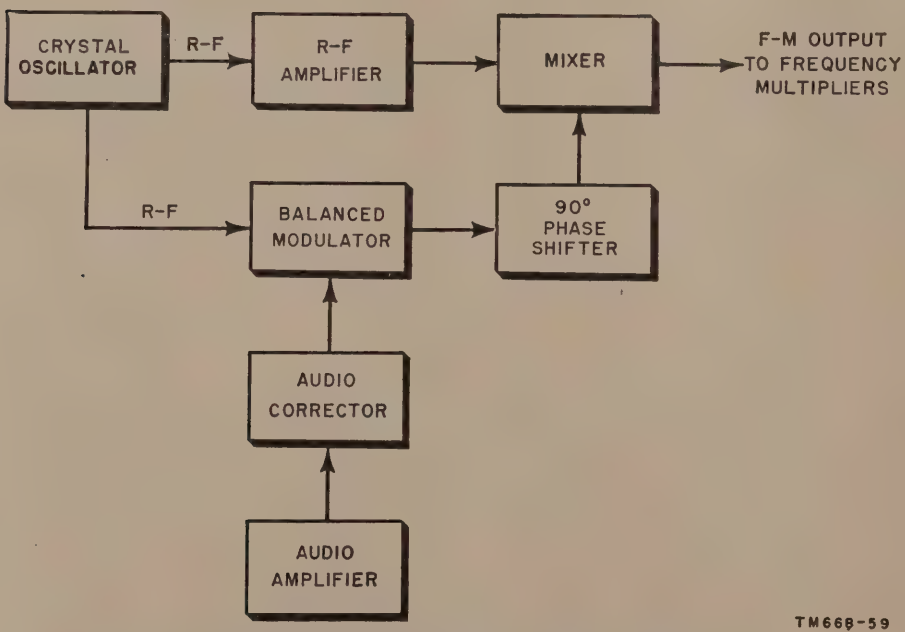


Figure 59. A-m to p-m to f-m—indirect method.

one of which is shown in figure 60. In this circuit the output from the crystal oscillator is coupled to the control grids of two tetrodes, from either side of a tuned transformer. Because of this method of coupling, the voltages at the grids of the tubes are 180° out of phase. The plates of these tubes are connected in parallel and an output is developed across a common load impedance, usually a parallel-resonant circuit.

plied to the screens of the two modulator tubes causing a variation in plate current which is proportional to the modulating voltage on each screen grid. The audio signal unbalances the modulator tubes and therefore the side bands produced by the audio and the carrier appear in the output. The situation can be understood in terms of the audio voltages applied to the screen grids. Since the input is positive on one control grid and negative

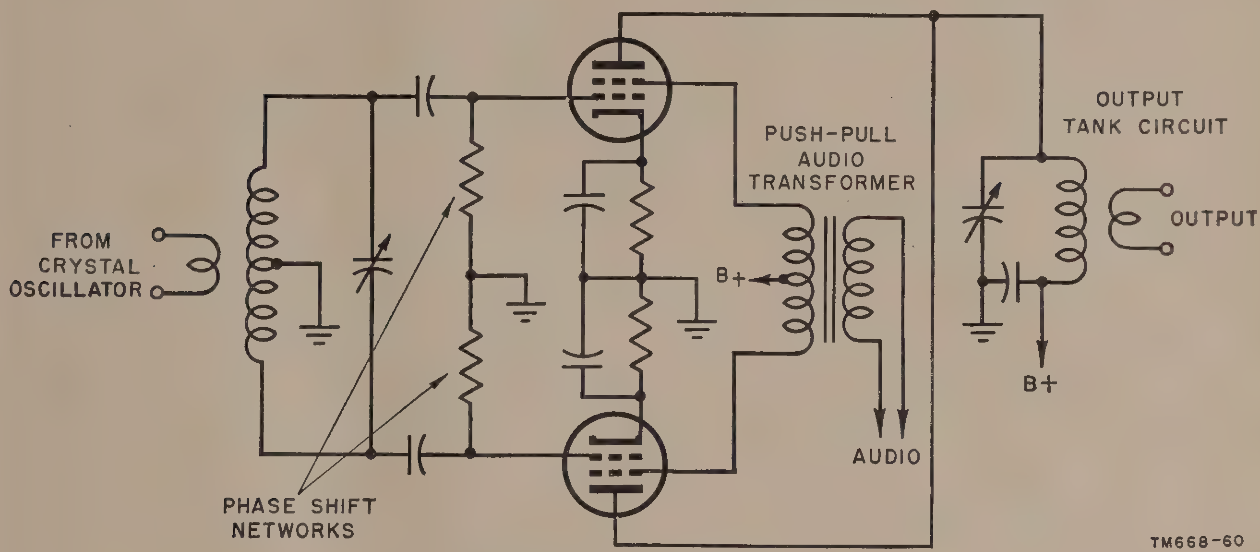


Figure 60. Balanced modulator.

(2) The grid excitation voltage passes through a resistor-capacitor network in which the reactance of the capacitor is much larger than the resistance. Therefore, the voltages at the grids, in addition to being out of phase with each other, are 90° out of phase with the voltage produced by the oscillator. If the two modulator tubes are identical, the variation of one grid voltage in a positive direction and the other in the negative direction produces equal and opposite changes in voltages at the two plates. Because the plates are connected in parallel, and the equal and opposite changes cancel, the input carrier is effectively eliminated across the common plate load. The voltage from the audio circuits is applied to the screens of the two modulator tubes causing a variation in plate current which is proportional to the modulating voltage on each screen grid. The audio signal unbalances the modulator tubes and therefore the side bands produced by the audio and the carrier appear in the output. The situation can be understood in terms of the audio voltages applied to the screen grids. Since the input is positive on one control grid and negative on the other, at any instant of time, and then reversed on the following half cycle, first one tube conducts and then the other. Consequently, output always will appear in the plate circuit from the tube which is conducting, whereas the other is almost at cut-off. When the screen of one tube is made more positive than the other by the audio voltage, the output increases. This action produces side bands equal to the sum and difference of the audio signal and the carrier, whereas the carrier is effectively canceled.

g. Other Balanced Modulator Circuits—Parallel-Grid Push-Pull Plate Circuit. It also is possible to produce the double side band without the carrier by connecting the control grids of the two tetrodes in parallel and the plates

in push-pull, applying the audio voltage to the screens, as before. With in-phase voltages on the control grids, equal voltages are produced at each plate. If these voltages are combined in a transformer, they are in phase at its opposite terminals. A transformer produces no output unless the voltages are out of phase on each side; therefore, the carrier cannot pass through the circuit. However, the voltages applied to the screens are in push-pull; that is, they are out of phase with each other. When the audio voltage is applied to the screen grids, the circuit is unbalanced, as before, and current flows in the output circuit. The 90° phase shift that is needed is incorporated in the output circuit by using a suitable network of coils and capacitors. It also is possible, by using the angle of lag in the current that flows in an untuned secondary, to make the output transformer itself act as a phase shifter.

h. Ring Modulator. A simple balanced modulator (fig. 61) that requires no vacuum tube can be constructed from four *varistors*. Varistors are special resistors made of material whose resistance *varies* with the applied voltage. In general, the resistance decreases as the voltage increases, so that the current flow increases at a rate much faster than in an ordinary resistor. The action is much like that of a simple vacuum-tube diode. Four copper-oxide rectifiers connected in a bridge circuit and sealed in a metal container commonly are called a varistor. The carrier is applied at two terminals, and the audio and load are applied to the other two. The carrier is balanced out in the bridge, and the unbalanced audio current that flows in the varistors produces the side bands in the output. These devices are used most widely in telephone work for the production of a double side-band pair, where they have the advantage of not requiring any currents besides those of the carrier and the modulating signal for their operation. If the double side band is recombined with the carrier, after a 90° phase shift, p-m is produced.

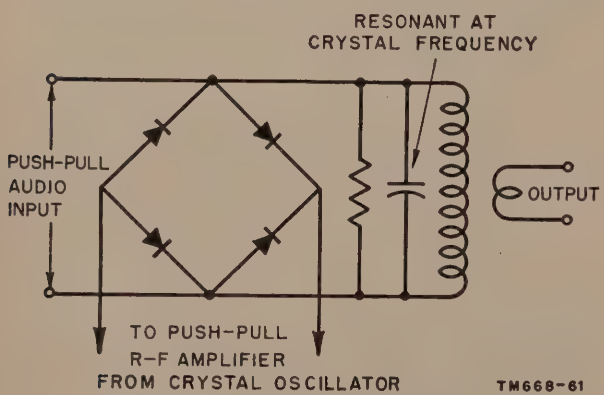


Figure 61. Varistor-ring modulator.

36. Modified Balanced Modulator

a. The balanced modulator and 90° phase-shift arrangement is satisfactory in terms of distortion. However, the deviation produced is very low, and a modification of the basic circuit has been worked out which does not require balancing out the carrier completely, shifting the side bands 90°, and reinserting the carrier. This circuit is used in many types of mobile and fixed equipment where simplicity and reliability are necessary. Basically, it is a cross between the Link modulator and the balanced modulator. A simplified version of this oscillator-modulator circuit is shown in A of figure 62. The grids of the pentagrid modulator tubes are excited through an R-C network in the output circuit of the oscillator. This network starts with a parallel-tuned circuit resonant at the crystal frequency. The opposite sides of the coil are 180° out of phase. This out-of-phase voltage is passed through a very small capacitor, *C*106, and the output is taken across a small resistor, *R*104. This introduces an additional 90° phase shift between one side of the coil and the other, for a total of 270°. The result is that the opposite grids of the modulator tubes are 270° out of phase. They are connected across a coupling network consisting of two capacitors, *C*104 and *C*107, and two resistors, *R*105 and *R*106. *C*105 serves to ground the center point between the two resistors. The modulator grids therefore are 45° out of phase in respect to the reference phase and 270° out of phase with each other.

b. In the vector diagram shown in B, the two 90° out-of-phase input signals, E_1 and E_2, are combined in the output of the modulator tubes to produce a resultant, designated as E_R. The two number 4 grids are fed by a push-pull audio signal. When the audio voltage on the grid of one tube goes negative, it reduces the transconductance of that tube. This, in turn, reduces r-f voltage and, correspondingly, the length of the

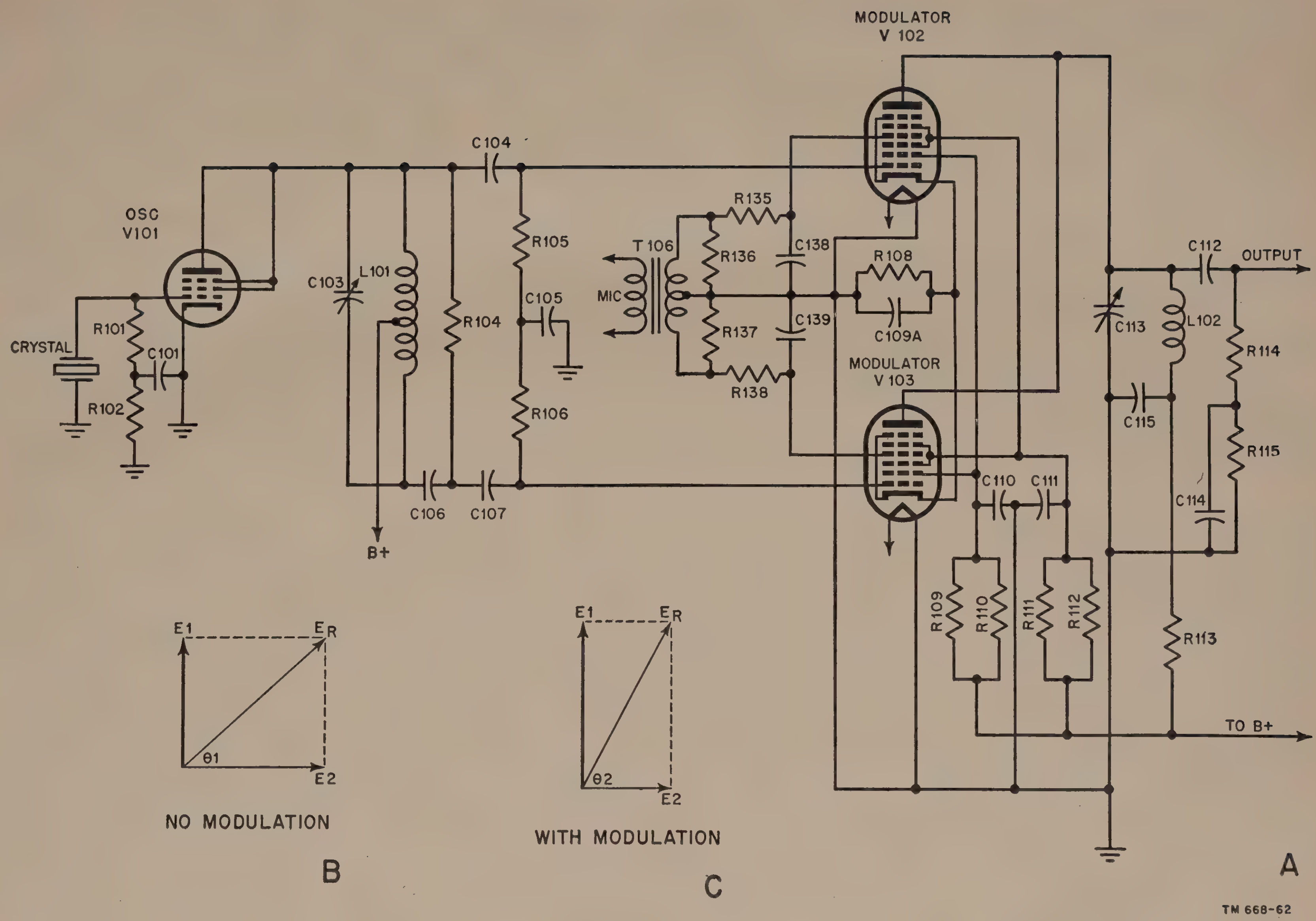


Figure 62. Modified balanced modulator.

vector representing the r-f voltage. Simultaneously, the other tube is being driven to a higher r-f output voltage by the positive part of the audio cycle applied to its grid. Therefore, the vector representing the increase in output from the second tube lengthens, as shown in C. The resultant of the shorter vector of plate voltage of one tube and the longer vector of the other swings from side to side, and a considerable amount of phase deviation is introduced. This deviation can never exceed the 90° difference in the r-f input signal to the two tubes. However, the distortion that is encountered usually dictates a lower value of phase deviation than indicated. The circuit resembles the balanced modulator very closely at first glance, but the difference between them lies in the fact that in this arrangement the grids are fed *90°* out of phase, and there is *no* cancellation of the carrier in the parallel plate circuit. In the balanced modulator, the grids are *180°* out of phase and the carrier *is* balanced out. Inherently, this circuit is not capable of the low distortion of the true balanced modulator carrier reinsertion arrangement.

Section III. SUMMARY AND REVIEW QUESTIONS

37. Summary

a. F-m always is produced near the frequency-determining section of the transmitter at a low power level.

b. Varying the inductance or capacitance in an oscillator in accordance with an audio signal produces f-m.

c. A reactance tube injects capacitance or inductive reactance into an oscillator circuit.

d. The transconductance of a vacuum tube is the ratio of a small change of plate current to a small change of grid voltage, with plate voltage held constant.

e. The amplification factor is the ratio of a small change in plate voltage to a small change in grid voltage, with the plate current held constant.

f. The plate resistance is the ratio of a small change in plate voltage to a small change in plate current, with the grid voltage held constant.

g. The product of the plate resistance and the transconductance equals the amplification factor.

h. The basic expression for the resistance and reactance injected by a reactance modulator is

$$Z = \frac{1}{g_m} + \frac{1}{g_m} \times \frac{Z_a}{Z_b}$$

where Z_a and Z_b are the impedances of the voltage divider connected between plate and grid, and between grid and cathode, respectively.

i. A reactance tube can inject capacitance or inductance, depending on the components of the voltage divider.

j. The input impedance of a tube changes with plate load or transconductance, and this change is known as the Miller effect.

k. A diode modulator acts as a variable resistor in series with a capacitor placed across a tank circuit, effectively changing the phase angle of the current drawn from it, and thus changing the frequency of the oscillator.

l. An R-C oscillator can be frequency-modulated by varying one of the frequency-determining resistors. Variation is accomplished by replacing a resistor with the dynamic plate resistance of a tube.

m. Crystal oscillators are more stable than conventional vacuum-tube oscillators, but they cannot be frequency-modulated directly.

n. A simple phase modulator introduces a variable time delay in a circuit, through which the carrier signal must pass. The variable time delay can be produced across one element of an R-C series circuit. The resistor can be replaced by the plate resistance of a tube.

o. A-m as well as p-m is produced in the simple phase modulator; therefore, the constant-impedance characteristic of a tuned circuit is used to overcome this disadvantage.

p. An R-C series circuit, with output taken across the capacitor, is used as an audio corrector to convert p-m to f-m.

q. The two ways by which a signal can reach the plate of a tube are grid-plate capacitance and tube transconductance.

r. The signal path, via the transconductance, can be varied by grid signal changes. This produces a variable voltage on the plate that combines with the out-of-phase capacitive voltage to produce a variable phase shift.

s. Greater phase shift can be obtained by feeding back part of the output into the grid circuit. An iron-cored coil in a tuned plate circuit also increases the effective deviation.

t. It is possible to produce phase modulation by varying the phase of a parallel resonant circuit by detuning it with the injected reactance of a reactance modulator. This parallel resonant circuit is used as the plate load in an r-f amplifier.

u. A nonlinear coil can be used to generate a series of phase-modulated pulses when both r-f and a-f currents flow through it at the same time.

v. When the modulation index is low, there is only a very slight difference between an f-m and an a-m signal.

w. The vector sum of the two side bands of a narrow-band f-m signal, or ordinary a-m signal, is known as the double side band. In f-m, the double side band is 90° out of phase with the carrier, whereas in a-m it is in phase with it.

x. A-m can be changed to p-m by removing the carrier, rotating the double side band 90°, and reinserting the carrier.

y. The double side band can be produced and the carrier rejected by a balanced modulator.

z. The grids of a balanced modulator are operated either in push-pull or in parallel, with the plates connected oppositely. The audio signal unbalances the modulator, permitting the side bands to go through.

aa. Varistors, arranged in a bridge circuit, can be used to produce a double side band without the carrier.

ab. In the modified balanced modulator, the grids are excited 90° out of phase. Considerably more deviation is obtained in comparison with the ordinary balanced modulator.

38. Review Questions

a. At what power level is f-m produced?

b. What varies the frequency of an oscillator?

c. What is the purpose of a reactance tube?

d. What are the three basic characteristics of a vacuum tube?

e. What is the relationship between them?

f. What are the four combinations possible for the voltage divider in the reactance modulator?

g. What is the Miller effect? Describe how it can be used to frequency-modulate an oscillator.

h. Describe the operation of a diode modulator.

i. How can an R-C oscillator be frequency-modulated?

j. What is the disadvantage of a conventional oscillator as compared with a crystal oscillator in an f-m transmitter?

k. What purpose does audio correction serve?

l. What happens to the phase of a parallel-resonant circuit, operating above its resonant frequency? Why?

m. Why must the amplitude of the r-f carrier, alone, be sufficient to produce saturation in the nonlinear coil modulator?

n. How are the phase-modulated pulses converted into continuous waves?

o. When is an f-m signal nearly equivalent to a-m?

p. What is the double side band?

q. What is the phase of the double side band, relative to the carrier, in f-m? In a-m?

r. How is an a-m signal converted into an f-m signal?

s. How are the input and output circuits of a balanced modulator connected?

t. How do the side bands pass through the balanced modulator?

u. Why is a 90° phase shift network put in the grid circuit of some balanced modulators?

CHAPTER 4
F-M TRANSMITTER

Section I. BASIC CIRCUITS

39. Basic Transmitter

a. The various direct and indirect methods for producing frequency modulation involve changing either the frequency or phase of an oscillator in accordance with some modulating signal. In the direct method, the modulating signal is injected into a modulator whose output varies the frequency of the oscillator in accordance with the original modulating signal. In the indirect method, the modulating signal is passed through a correction network to a phase modulator. The correction network changes the phase of the modulation in such a manner that, when the output of a crystal oscillator is passed through the modulator, the oscillations are frequency-modulated in accordance with the modulating signal.

b. It is extremely important that the transmitter be transmitting at its designated frequency. To achieve maximum frequency stability, therefore, the oscillator is operated at relatively low frequencies, and the lower the frequency, the more stable the oscillator. This means that the center frequency of the f-m signal output of the modulator-oscillator section is lower than the carrier frequency desired for transmission. To raise the f-m signal to the correct frequency, it is passed through a series of *frequency multipliers.* Each stage of frequency multiplication raises the frequency of the signal input by some multiple of the fundamental frequency: A doubler produces a signal at its output that is twice the frequency of the input signal; a tripler raises the frequency three times; a quadrupler four times. When the input to the frequency multiplier is an f-m signal, the multiplier produces an increase in the frequency deviation.

c. The oscillator-modulator and frequency-multiplier sections of the transmitter are operated at low power levels, and the output of the final multiplier is too weak to be transmitted. A power amplifier similar to those in a-m transmitters acts as the final stage in the f-m transmitter to build up the signal to the power level desired.

d. The indirect method of f-m transmission generally uses a crystal oscillator to produce the r-f signal to be modulated since the crystal increases the frequency stability. In direct f-m, a crystal oscillator with its fixed frequency cannot be used, because the oscillator must be free to change frequency in accordance with the modulating signal. Whatever the type of direct f-m oscillator, it cannot be made as stable as a crystal oscillator, and an auxiliary circuit must keep it on the correct center frequency. Such a circuit is called *afc,* or *automatic frequency control.* A complete description of the f-m transmitter covers all the circuits mentioned above. Regardless of the circuits used in a particular transmitter, however, it must be remembered always that the *sole* purpose of the equipment is the transmission of intelligence from one point to another.

40. Frequency Multiplication

a. General. In f-m transmitters, frequency multiplication of the f-m signal performs two functions: It increases the frequency of the signal to the value desired for transmission, in this way acting the same as a frequency multiplier in an a-m transmitter. It also increases the effective frequency deviation of the f-m signal.

b. *Increasing Frequency Deviation.*

(1) The f-m signal from the oscillator-modulator section has a center frequency f_c and a frequency deviation of $\triangle f$ caused by the modulating signal. The f-m signal therefore varies from a maximum of $f_c + \triangle f$ to a minimum of $f_c - \triangle f$. For example, with an oscillator whose unmodulated output frequency is 100 kilocycles, a certain audio signal causes this frequency to swing between 95 and 105 kc, and the frequency deviation therefore is ± 5 kc.

(2) If this f-m signal is impressed on the grid of a tube which is operating as a doubler, the center frequency of the doubler output is twice f_c, or 200 kc. Since the multiplier doubles any frequency appearing at the grid, when the f-m signal is deviated to 95 kc, the output frequency is 190 kc. When the f-m signal is 105 kc, the output is 210 kc. The multiplier output therefore varies from 190 to 210 kc, and the new deviation is 10 kc. By doubling the frequency at its input, the multiplier also has doubled the frequency deviation. The amount of multiplication used depends on the frequency to which the signal must be raised and the amount of frequency deviation desired. The greater the deviation, the greater is the bandwidth of the f-m signal transmitted.

c. *Basic Frequency Multiplier.*

(1) Essentially, frequency multipliers are harmonic generators; that is, the output frequency is some multiple of the input frequency. The output circuit must contain not only the original input frequency, but also harmonics of it, and is made selective to the harmonic desired, all other frequencies being rejected. To produce these harmonics, a class-C amplifier stage is used with its plate circuit tuned to the frequency of the harmonic desired. Such an amplifier is called a frequency multiplier.

(2) Figure 63 is the simplified schematic of a frequency doubler using a tube operated in class C. The input signal is impressed across the grid circuit through transformer *T*1. The frequency of this signal is the fundamental frequency, *f*, of the system. Because of class C operation, the plate current is nonlinear and therefore rich in harmonics. If a tuned circuit, *L*2-*C*2, is placed in the plate circuit, it can be tuned to the frequency of the desired harmonic. When *L*2-*C*2 is tuned to twice the frequency of *L*1-*C*1, this stage becomes a doubler.

(3) In figure 64, a typical i_b-e_c characteristic curve for a tube in class C operation is shown. The grid is biased far below cut-off, and plate current flows only during the portion of the signal that takes the grid voltage above cut-off. This portion of the signal is only a fraction of the positive half-cycle, and the resultant plate-current flow is in the form of short pulses during the

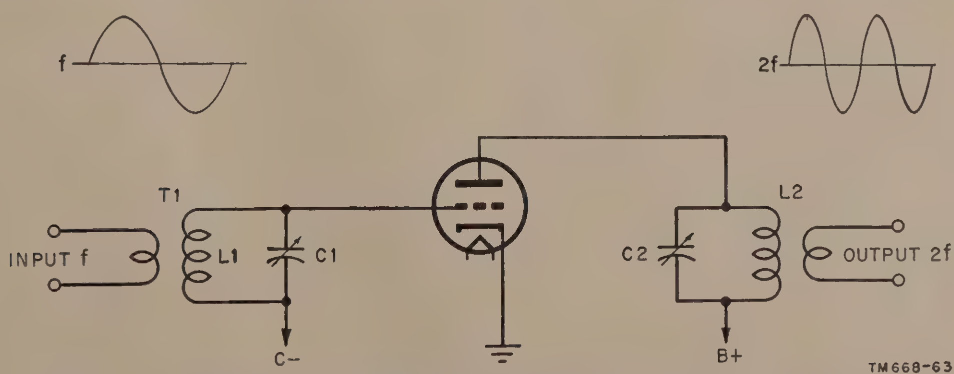


Figure 63. Simplified schematic of frequency doubler.

time the grid voltage is above cut-off. The amplitude of these pulses depends on the amplitude of the signal at the grid.

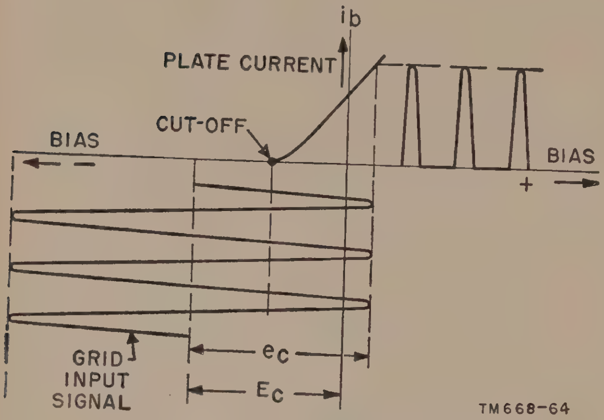


Figure 64. Class C amplifier operation.

(4) The output pulses of plate current are distinctly nonlinear in respect to the input signal, and it is this characteristic of class C operation that is used to produce an output at the harmonic frequency. If the output waveform exactly reproduced the sinusoidal input waveform, then no frequencies could be present in the output that were not present in the input. If the *shape* of the waveform is changed, new frequencies must have been added. These new frequencies take the form of harmonics of the input frequency. The frequencies present depend on the duration of the pulse and how sharply it is peaked. The shorter and sharper the pulse, the more harmonics are produced in the output. With a doubler, for example, the tube is biased farther beyond cut-off than in a class C amplifier. This produces a shorter and sharper pulse which contains sufficient energy at the second harmonic to drive the tank circuit. Biasing the tube farther beyond cut-off requires a larger input signal to produce the same amplitude of current flow in the plate circuit.

(5) The amount of multiplication depends on *the final frequency and frequency deviation desired.* A single multiplier can be used to multiply the frequency by as much as five times. The higher the order of multiplication, however, the lower the output of the stage. Usually, the desired multiplication is obtained through several successive stages, the highest-order multiplier used being the quadrupler. For example, the center frequency of the signal from a modulator-oscillator section is 10 megacycles, and it is desired to transmit this signal at a center frequency of 80 mc. This means multiplying the signal frequency eight times. Two possible ways of obtaining this amount of multiplication are shown in figure 65. In A, three multiplication stages are used. Each stage is a doubler with its output tank tuned to the second harmonic of the signal input at its grid. The first stage raises the center frequency of the signal from 10 to 20 mc. This is applied to the center frequency of the input of the second doubler, which raises the center frequency to 40 mc. The last doubler stage raises the signal to the desired center frequency of 80 mc.

(6) In B, the same result is obtained by using a frequency quadrupler followed by a doubler. The frequency is multiplied four times in the first stage, and the output then is fed to a doubler. This produces the same amount of multiplication as the three doublers used in the first method, and with only two tubes. The second method is used where compactness and economy are desirable, but efficiency and power output are lower.

d. Push-Push Doublers. When the plate tank of the circuit in figure 66 is tuned to twice the frequency of the input, the circuit acts as a very efficient doubler. Its operation can be considered similar in action to that of a full-wave rectifier. The grid of each tube is biased approximately to cut-off so that plate current flows in each tube on succeeding half-cycles. When the signal input across the secondary of *T* makes the grid of *V*1 positive in respect to its cathode, the tube conducts; at the same time,

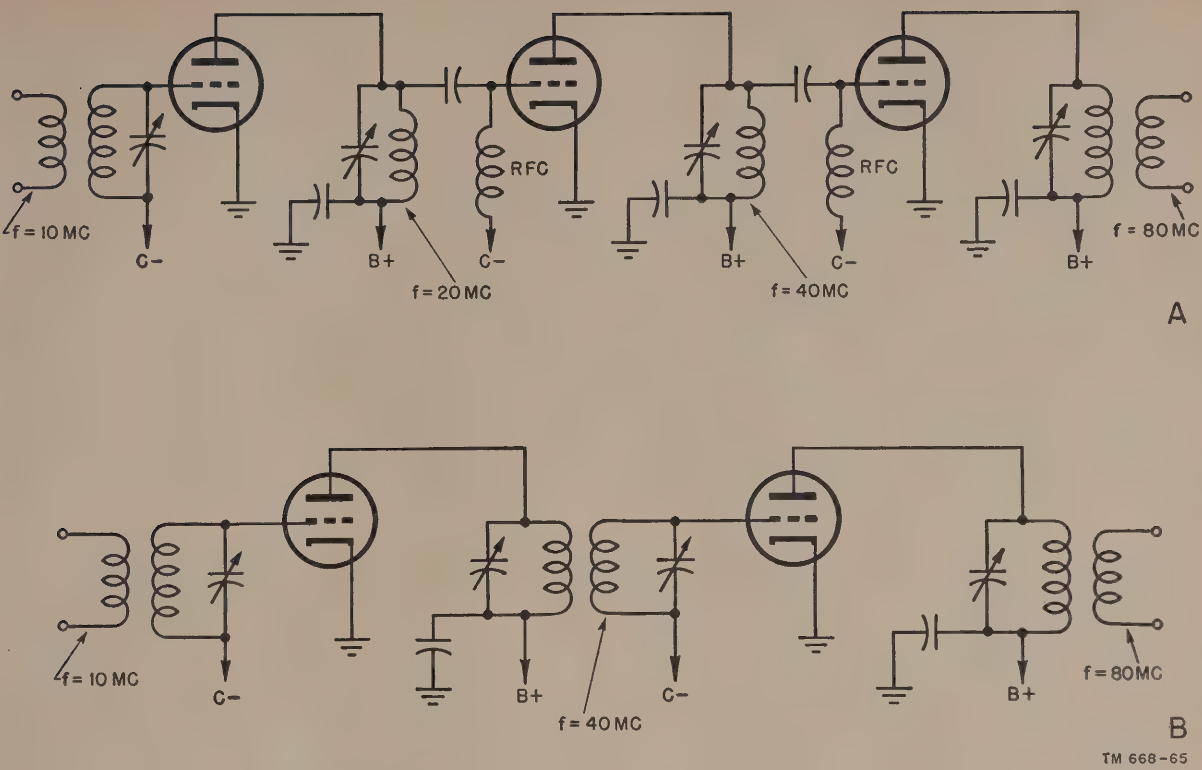


Figure 65. Two typical frequency-multiplier chains.

the signal applied to the grid is of $V2$ negative and $V2$ remains cut off. On the next half-cycle of input voltage, $V2$ conducts and $V1$ is cut off. The plates of $V1$ and $V2$ are connected in parallel; therefore two pulses excite the tank circuit for each cycle of input. These pulses therefore drive the tank at a frequency twice that of the input. The output can be compared to the ripple present in the output of the unfiltered full-wave rectifier circuit. Because of its simplicity, this circuit is widely used in f-m transmitters. If compact design is desired, the circuit can serve also as combination doubler and power amplifier because of its relatively high efficiency and low output of undesirable harmonics.

e. Multiplier Operation at High Frequencies. The operation of frequency multipliers at high frequencies is hindered by degenerative effects tending to lower the power output. These effects can be caused by capacitive or inductive feedback. Capacitive coupling between grid and plate circuits caused by the interelectrode capacitance gives rise to degeneration which is equivalent to loading the output circuit. It has

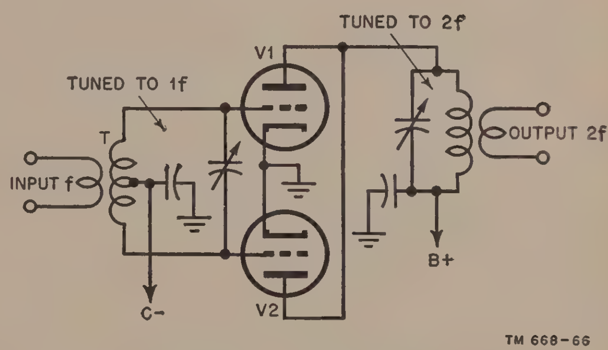


Figure 66. Push-push doubler.

been found that power tubes with large values of mutual conductance and grid-plate capacitance are rendered inoperative as multipliers at high frequencies by this loading effect. Before the circuit can be made to operate with full efficiency and power output this loading effect must be *neutralized*. Three methods of canceling the effects of the grid-plate capacitance are illustrated in figure 67.

(1) The circuit in A shows one method of neutralizing the grid-plate capacitance, C_{gp}. Capacitor C_n provides feed-

back which is 180° out of phase with the degenerative feedback caused by C_{gp}. By selecting the correct value of C_n, the two feedbacks are made equal, but out of phase, and cancel. Care must be taken not to make C_n too large, or the regenerative feedback will cause oscillation. C_n usually is adjusted for minimum d-c plate current with the plate circuit at resonance. Another method, shown in B, is to insert from plate to grid a series *L-C* circuit which is antiresonant at both input and output frequencies. This has the effect of placing a high-impedance feedback path from plate to grid. C_n is adjusted for minimum plate current with the circuit at resonance. If the input circuit between grid and cathode presents a low inductive reactance to the feedback voltage, it will serve to balance out the effects of degenerative feedback. By inserting the proper amount of inductance, as shown in C, in series with the grid circuit, the degenerative feedback is canceled. L_n must not be made too large, or the circuit will oscillate. In a v-h-f (very-high-frequency) multiplier, lengthening the grid lead may be sufficient to provide this inductance.

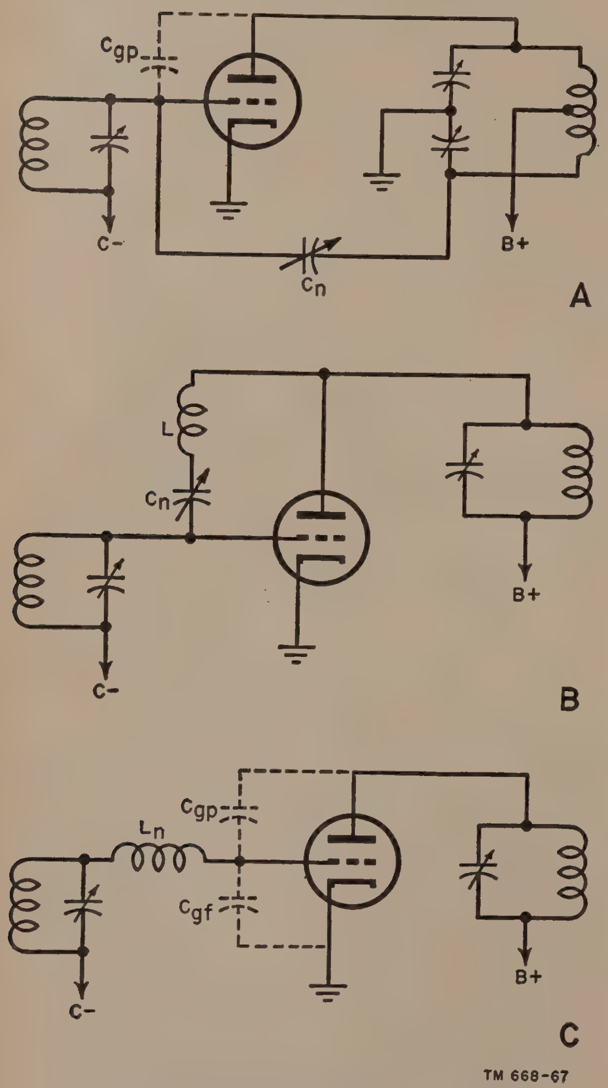


Figure 67. Neutralizing circuits.

(2) Degeneration equivalent to an output loading effect also can be produced by cathode inductance common to both plate and cathode circuits. This inductance prevents the cathode from being connected directly to ground, and plate current flowing through the cathode circuit creates a reactive voltage which opposes the input signal. To obtain efficient operation at very-high frequencies, the cathode inductance must be neutralized at both input and output frequencies. A method for accomplishing this is illustrated in figure 68. The circuit is a series combination of *L*1 and *C*1 shunted by capacitor *C*2, the combination being in series with the cathode lead. *L*1 and *C*1 in series with L_k form a circuit providing series resonance at the input frequency. L_k and shunt capacitor *C*2 form a series resonant circuit to ground at the output frequency.

(3) Several tetrodes and pentodes have been designed especially for use at very-high frequencies. In these, two separate leads for grid and plate returns to the cathodes have the effect of reducing degeneration. When high output and efficiency are needed, these tubes can be used in a push-pull arrangement. Also, if the screen grid of a tetrode or pentode is effective at the frequency of operation, no neutralization is required.

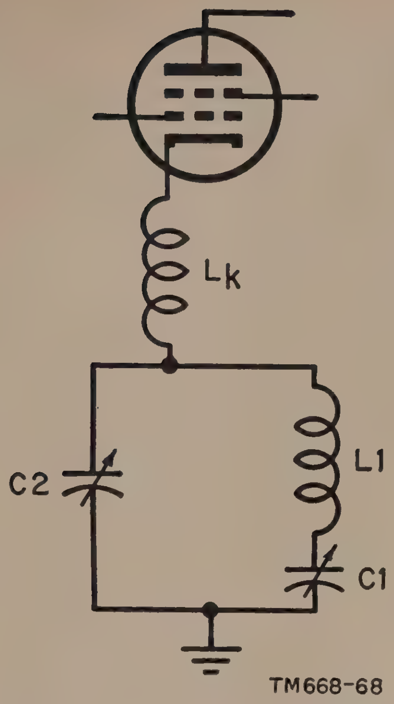


Figure 68. Cathode inductance neutralization.

f. Other Frequency Multipliers.

(1) Mixers similar to those in superheterodyne receivers can be used for frequency multiplication. However, they give no increase in effective deviation. It is possible also to synchronize an oscillator running at a higher frequency with one at a lower frequency, if the two frequencies are multiples of one another. The synchronized oscillator is used more frequently, however, as a frequency *divider.*

(2) Higher multiples of a given frequency can be obtained by using a nonlinear device that produces harmonics. The distortion of the grid current in a class C amplifier is one method. A second method produces harmonics in a mixer or other nonlinear modulator. The desired harmonic is amplified and then fed back to the mixer, where it reinforces the output at its own frequency. This device is called a regenerative modulator. Since subharmonics can be selected as well as harmonics, the device can be used also as a frequency divider. In practice, this has been its principal application.

g. Combined Frequency Multiplier, Master Oscillator.

(1) The master oscillator and a multiplier can be combined in a circuit using only one tube. In figure 69, such a circuit combines the oscillator and multiplier in a single pentode tube. Elimination of components and reduced current drain are gained at the expense of only a slight loss in oscillator stability.

(2) The oscillator is a simple Colpitts, with r-f oscillations generated in the control-grid, screen-grid circuit of the tube. The output frequency is selected by a tuned plate load, *L*2-*C*2. The tank circuit, *L*1-*C*1, in conjunction with *C*8 and *C*9 forms the fundamental frequency-determining components. The grid-leak bias for the operation of the oscillator is provided by *R*1 and *C*4. The r-f choke in the cathode circuit permits d-c current to re-

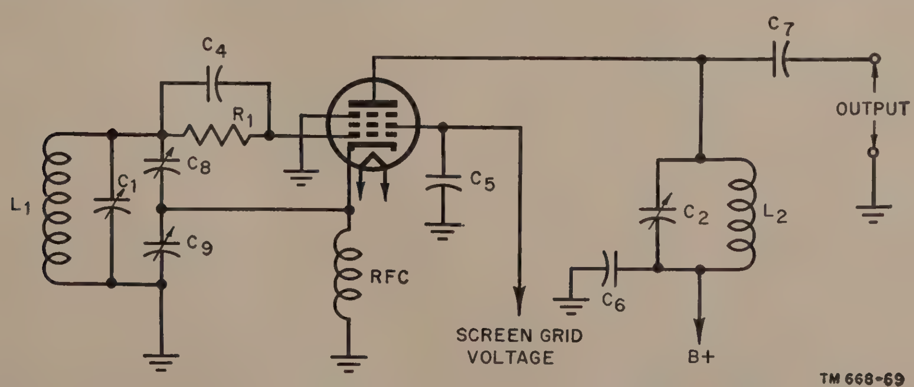


Figure 69. Typical combined oscillator, frequency multiplier.

turn to the negative side of the power supply, at the same time maintaining the r-f potential at the cathode. The screen is effectively bypassed for r-f by *C*5. Therefore, the screen can act as a grounded plate, with the control grid and cathode serving their normal functions. The result is a triode oscillator circuit.

(3) The current pulses generated by the oscillator section reach the plate flow through the tuned resonant circuit formed by *L*2 and *C*2. This circuit presents a high impedance to the harmonic frequency of the plate current pulses since it is tuned to resonance with it. Consequently, a considerable harmonic voltage is developed between plate and ground. The lower part of the plate tank circuit is bypassed to ground through capacitor *C*6 and the output from the stage is coupled capacitively to the following stage through *C*7.

(4) Since the output circuit is coupled to the oscillator circuit through the electron stream alone, there is comparatively little interaction. If the screen voltage is set properly, it is possible to reduce the variation in operating frequency with changes in tuning of the output circuit to a low value. The higher the order of harmonic to which the plate circuit is tuned, the better the stability of the oscillator. Any interaction between output and oscillator circuits must come as a result of Miller effect between the two circuits. Capacitive coupling between the grid and plate, which tends to cause this interaction, is considerably reduced by the shielding effect of the screen grid. Therefore, the grounding of this grid through capacitor *C*5 must be complete to obtain maximum isolation.

(5) The circuit of figure 69 is one possible way in which a frequency multiplier can be combined directly with the oscillator. Any oscillator circuit that can operate with its plate at ground potential can be substituted for the Colpitts circuit shown. The frequency-multiplier action is the same regardless of the oscillator, the only advantage gained with any specific circuit being attributable to the characteristics of the oscillator itself.

41. Power Amplifiers

a. F-M Power Amplifiers.

(1) The requirements for an f-m power amplifier are somewhat different from those for a-m, in which the power amplifier is usually the stage in which the modulation is introduced. Therefore, any losses that take place during the modulation process must be dissipated in the power amplifier stage. Since the f-m power amplifier has no connection with the modulation process, the only losses that are involved are those inherent in the tube and circuit when amplifying an unmodulated carrier.

(2) When a-m is produced in a low-level stage, it is necessary that the power amplifiers reproduce the modulation envelope without distortion, and therefore linear amplifiers must be used. F-m, which is also produced at a low level, does not have a modulation envelope that can be distorted by the limiting action of highly efficient class C amplifiers. The characteristics of an f-m power amplifier are determined at class C, c-w ratings.

b. Class C Amplifiers.

(1) A typical class C amplifier, as used for f-m signals, is shown in figure 70. The input signal is supplied through a tuned transformer, *T*1. Output is developed by the r-f signal appearing across the parallel-resonant circuit, *T*2.

(2) Figure 71 shows the relationship between the various voltages and currents in the circuit of figure 70. The grid bias, e_c, is developed across the tuned circuit of *T*1. The class C amplifier operates with grid bias much greater than cut-off. Therefore, the

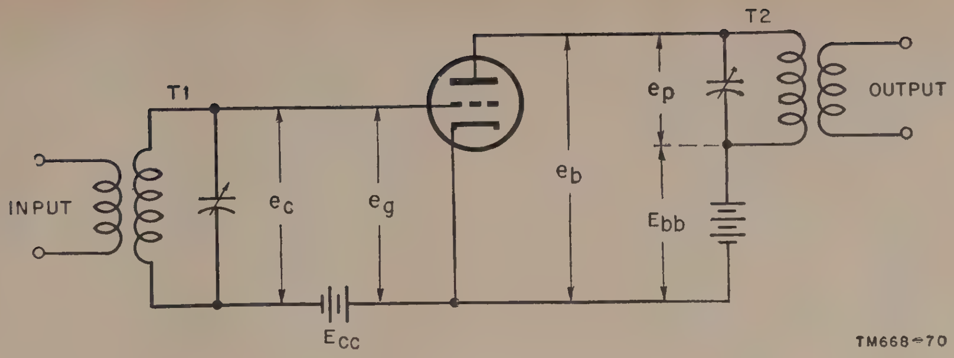


Figure 70. Class C amplifier.

grid excitation voltage causes plate current to flow during only part of the cycle. During the remainder of the cycle, the voltage on the grid is below the cut-off value, the plate current, i_b, is zero, and the corresponding plate voltage, e_b, rises to its highest value, or E_{bb}. Since no plate current flows, the voltage drop across the plate load impedance must be zero. The voltage drop across the load, therefore, is 180° out of phase with the grid voltage. The a-c components of the plate and grid voltages are sinusoidal because of the sharply tuned resonant circuits.

(3) Plate current flows when the grid voltage, e_c, rises above cut-off. The angle of flow of plate current is ϕ_p and is usually less than half a cycle. Grid current flows during the angle ϕ_g when the grid voltage, e_c, becomes positive. The sum of these two currents, $i_b + i_c$, is the space current, i_s, and represents the total current leaving the cathode. The angle of grid current flow depends on the *ratio* of the *grid bias* to the *peak signal amplitude*. This is equivalent to saying that, in a particular amplifier, the value of the grid bias chosen determines the angle of plate current flow for a given input signal. Short angles of flow give high efficiency and low power output, whereas large angles give low efficiency and higher power output.

(4) At any moment, the total power input to the plate is the product of the total voltage, e_b, supplied to the plate, and the instantaneous plate current, i_b. The power output is equal to the product of the load voltage and the plate current. The power loss at the plate is the difference between the input power and the output power. The efficiency of a class C amplifier is the ratio in percent of the output to input power and is usually between 60 and 80 percent. This high efficiency is possible because the plate current flows only when most of the voltage drop is across the output circuit. Therefore, only a small part of the supply voltage is wasted as a voltage drop between the plate and cathode of the tube.

(5) Since the grid of the tube swings positive and draws current during part of the cycle, power is absorbed from the excitation circuit, which is the product of the exciting voltage, e_c, and the grid current, i_c. Some of this power is lost at the grid, and the remainder is dissipated in the bias battery. If grid-leak bias is used, the remainder is dissipated as heat in the grid-leak resistor.

c. Class C Power Amplifiers With F-M Excitation.

(1) An f-m wave will not be distorted in passing through such an amplifier, since the frequency of the voltage developed at the output is the same as that provided by the grid excitation. If the input signal deviates through

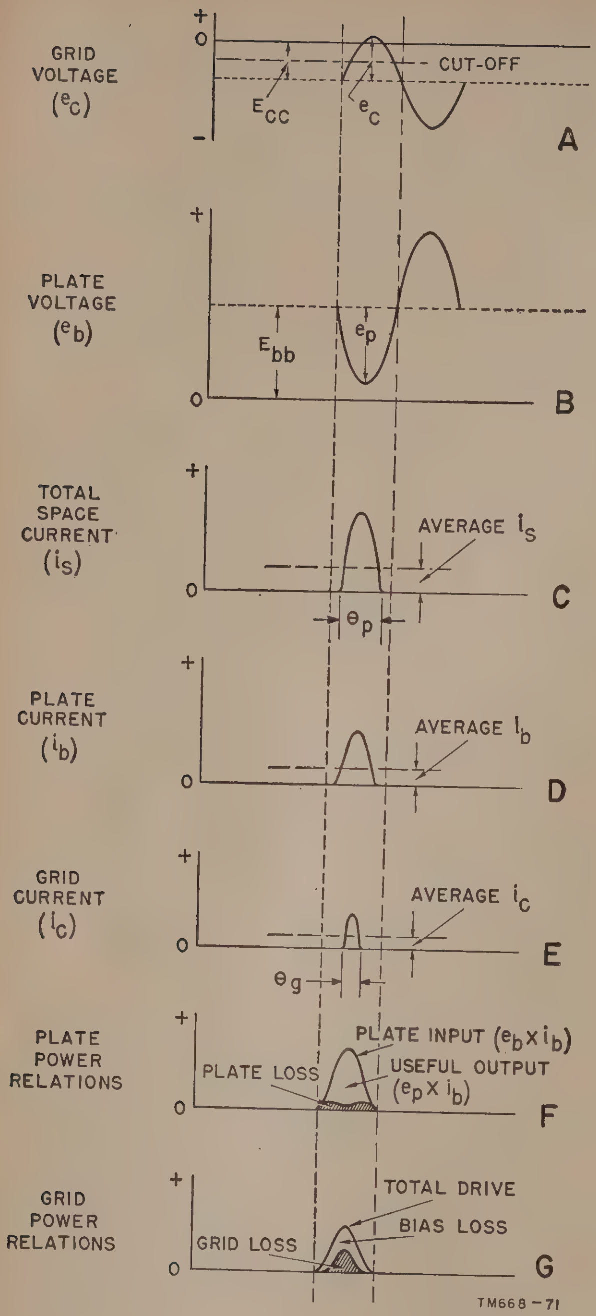


Figure 71. Current and voltage in Class C amplifiers.

a number of cycles, the output signal deviates by the same amount. Since resonant circuits are used in both input and output circuits, it is necessary that they have broad enough selectivity that the full frequency band of the modulated carrier is passed without reducing the amplitudes of the outer sidebands. This is possible only where the total bandwidth of the f-m signal is not large compared with the carrier frequency. This places a restriction on the maximum usable deviation.

(2) An examination of the circuit in figure 70 shows that, since the grid and plate circuits are tuned to the same frequency, energy can feed back through the grid-plate capacitance, permitting tuned-plate, tuned-grid oscillation at the frequency of the tuned circuits. In all triode amplifiers, this tendency toward oscillation must be neutralized. Figure 72 shows the same circuit as figure 70, with the addition of a tapped plate-tank coil grounded at the center by a bypass capacitor. From the side of the coil opposite the plate connection, a small variable capacitor, C_n, is connected to the grid. Since the opposite sides of the coil are 180° out of phase, the voltage tapped by the small variable capacitor is therefore 180° out of phase with the plate voltage. It is also out of phase with any grid voltage fed back through the grid-plate capacitance. The variable capacitor, along with the interelectrode capacitance from grid to cathode (shown in dashed lines), acts as an adjustable voltage divider which permits a variable amount of out-of-phase voltage to be applied to the grid. The voltage tending to cause oscillation is out of phase with this neutralizing voltage. It therefore can be canceled out if the two are made equal by proper adjustment of the neutralizing capacitor. The neutralizing capacitor is of approximately the same size as the grid-plate capacitance. It usually is made slightly larger in practical high-frequency circuits. This is done because inductance in the connecting leads produces reactance opposite to that of the capacitor and tends to reduce its effectiveness.

d. Class C Tetrode Amplifiers.

(1) The disadvantages of ordinary triode amplifiers in the frequency range

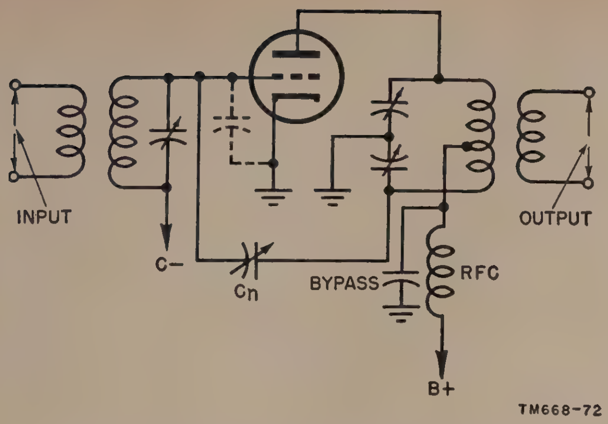


Figure 72. Neutralized triode amplifier.

where f-m usually operates are many. The necessity for neutralization means that an additional adjustment is needed and, as the frequency increases, this becomes more and more critical. Because of their low power sensitivity and high output capacitance, a large amount of grid excitation power is needed to produce a given amount of output power and the excitation requirements increase as the frequency increases. The high output capacitance of triodes also reduces the value of usable tank-circuit inductance, resulting in high Q and too narrow a band pass. At higher frequencies, where it is important to conserve the number of tubes and the amount of total power input, tetrodes are more useful.

(2) The operation of a tetrode amplifier for the very-high frequencies used for f-m transmitters is somewhat different from that for the triode. The grid bias is set with reference to screen-current cut-off, as compared to plate-current cut-off in a triode. The angle of flow therefore depends largely on the screen-grid voltage and control-grid bias. The maximum grid voltage must not be greater than the screen voltage. In addition, the minimum plate voltage during the operating cycle must not be less than the screen voltage (except in beam tetrodes). If this last condition occurs, the plate emits secondary electrons that are collected by the screen, making the plate a virtual cathode. The space charge in a beam tetrode (or the suppressor in a pentode) prevents this effect.

(3) In the higher part of the v-h-f range, it is no longer possible to use many tetrodes as class C amplifiers for f-m. At these frequencies, the inductance of lead wires and internal tube supports produces enough reactance to prevent the screen from being effectively grounded to r-f currents. Therefore oscillation can occur. For limited frequency ranges, it is possible to reduce the impedance from screen to ground by making the screen bypass capacitor and the screen lead inductance a series resonant circuit. However, if the circuit must be used over a wide frequency range, which is common with f-m equipment, an additional screen neutralizing control must be added, the operation of which is very critical. To overcome this difficulty, special circuits must be used.

e. Grounded-Grid Triode Amplifiers.

(1) There are three types of triode amplifiers, the type depending on the manner in which the signal is applied to obtain an output. The load may be connected between the plate and cathode, with the signal applied between the grid and cathode. If the common element is placed at zero potential, the stage is called a *grounded-cathode* amplifier. In the *cathode follower,* the plate is grounded to r-f, the signal is applied between grid and ground, and the load is placed between the cathode and ground. Finally, in the *grounded grid amplifier* (fig. 73), the signal is applied between the cathode and ground, the grid is grounded, and the output is taken across a load between plate and ground.

(2) The grounded-grid circuit permits a triode to be operated at high frequen-

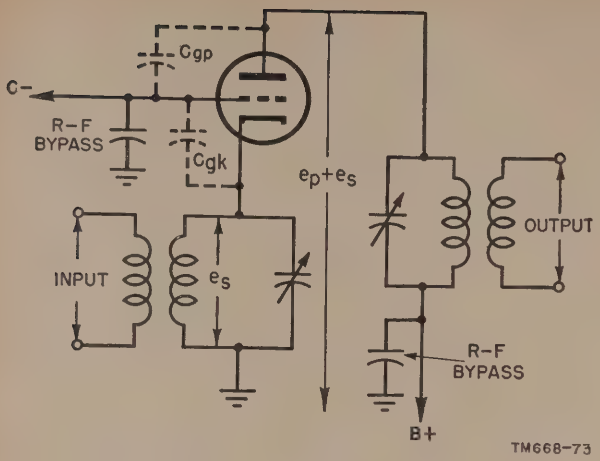


Figure 73. Grounded-grid amplifier.

cies without neutralization. Therefore, one of the most objectionable features of a triode r-f power amplifier is overcome. In this circuit, the grid is grounded through an r-f bypass capacitor and serves as a shield between the input and output circuits, thus preventing feedback of energy and resultant oscillation. It also has the advantage of very low output capacitance, since the only capacitance across the output added by the tube is that between grid and plate (fig. 73). In tubes designed especially for this purpose, the capacitance is made very low and larger values of inductance can be used in the plate circuit at relatively high frequencies. This results in higher efficiency.

(3) Another characteristic feature of the grounded-grid amplifier is that both the driver stage, which supplies the input, and the amplifier stage itself supply the plate load circuit. Note that the driver produces an r-f voltage e_s across the input terminals. An r-f voltage also is produced across the plate and cathode elements of the tube. These voltages are 180° out of phase in respect to the cathode, and therefore the r-f output voltage from plate to ground is the sum of the two out-of-phase voltages.

(4) The plate current generally is 180° out of phase with the plate voltage. This means that the signal current flowing in the cathode circuit must be the same as the plate current. The cathode can have low impedance, and the plate circuit can have high impedance. Therefore, the tube acts as a device to transfer the space current from a low impedance to a high impedance. The output power is proportional to the square of the current multiplied by the resistance; therefore, the input (cathode power) is low, the output (plate power) is high, and the tube acts as a power amplifier. The gain of the amplifier is proportional to the ratio of the output impedance to the input impedance.

(5) Because the input impedance can be made small, the bandwidth of the input circuit can be very great. Since the output capacitance is only that from plate to grid, the inductance in the plate circuit can be made large for a given resonant frequency. Therefore, the selectivity of the output circuit also can be made broad.

f. Push-Pull Amplifiers for F-M.

(1) The grounded grid push-pull amplifier is used frequently in high-power f-m systems at very-high frequencies. This circuit can be used also in the microwave region with specially constructed tubes and circuits. When two tubes are within one envelope, degeneration resulting from the inductance of the cathode lead is canceled out. This permits the use of lower values of grid drive than would be necessary with two separate tubes or with a single-ended stage. Typical circuits for push-pull triode and beam-tetrode amplifiers are shown in figure 74. Because two tubes are involved, twice the amount of excitation must be supplied. However, the output capacitance placed across the plate tank circuit and the input capacitance across the grid tank are halved because the tube capacitances are in series, and therefore equal to only half the value of one alone.

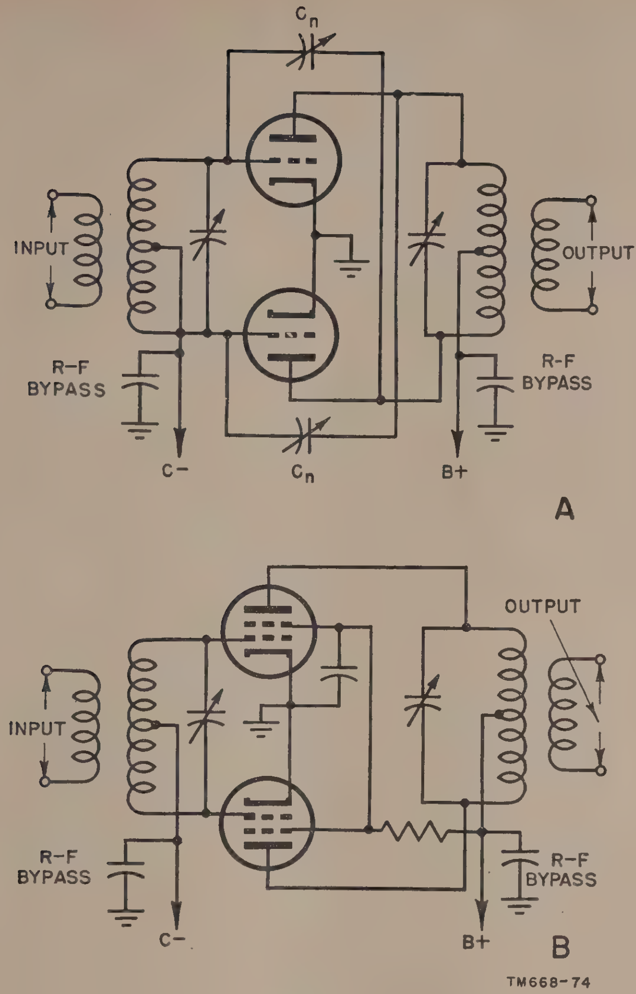


Figure 74. Push-pull amplifiers.

(2) At high frequencies, the necessity for neutralization of the push-pull triode amplifiers with its attendant difficulties makes the use of tetrodes desirable. These are combined in one envelope for high frequencies so that the internal inductance of the leads and tube elements does not interfere with operation. The usual tube construction has a single cathode which eliminates the problem of separate cathode-lead inductance. Neutralization of any amplifier increases the output capacitance of the tube. The push-pull cross-neutralized amplifier shown in A of figure 74 has an output capacitance equal to the plate-cathode capacitance of each tube in series plus twice the grid-plate capacitance. This extra capacitance is added by the neutralizing circuit. It limits the operating frequency of the amplifier because the output capacitance is a major factor in determining the plate-tank constants at very high frequencies.

(3) Some of these difficulties can be overcome by using push-pull, grounded-grid amplifiers. The operation of these amplifiers is similar to the single-ended stage, the major difference being the change in the cathode input impedance. Both cathode loads are effectively connected in series, and the input impedance becomes four times the value of one tube alone. Each part of the cathode load acts independently for its associated tube and therefore the voltage across both loads is twice that of the individual load. When the voltage across the secondary is doubled, the impedance, which is proportional to the square of the voltage, is increased by a factor of four. The same thing is true of the output load impedance in all push-pull amplifiers. The voltage across each section of the primary is the same as for a single-ended amplifier. The doubled voltage of the push-pull connection requires a total load impedance of four times the impedance for one side of the load. This makes the requirements for tank-circuit inductance easier to meet at very-high frequencies and accounts for the wide use of push-pull circuits for f-m power amplifiers.

g. Power-Amplifier Input Circuits.

(1) It is highly desirable to have as efficient a transfer of power from the driver stage to the power amplifier as possible. Therefore, the grid tank circuit must provide an impedance match between the grid input impedance of the amplifier and the plate output impedance of the driver stage. If a grounded-grid amplifier is used, similar considerations apply to the cathode tank circuit.

(2) The impedance of a circuit normally is defined as the ratio of voltage to current. However, in the grid circuit of a class C amplifier, this ratio is far from constant. When the grid voltage goes highly negative, no current is drawn at all; when it is positive, a great deal of current flows. Therefore, the impedance of the grid circuit varies over a range from an extremely high to an extremely low value through the operating cycle. If the input impedance of the grid circuit is too high, the heavy current demanded by the extreme positive grid swing cannot be drawn. As a result, actual grid voltage and consequent loss of peak efficiency are reduced in the operation of the amplifier. If the impedance of the grid tank circuit is too low, a great deal of power from the driver stage is required to operate it, and the losses in the inductor consume a considerable amount of the applied power. Generally, a compromise value is used which is approximately equal to the ratio of the driving power in watts divided by the square of the grid current. The choice of values for the components in the grid tank circuit is determined by this impedance. The result usually is satisfactory regulation of the grid voltage without excessive power loss.

(3) Some of the actual circuits used at the grids of single-ended, push-pull, and grounded-grid amplifiers are shown in figure 75. The simplest of the capacitance-coupled, tuned-input circuits for a single-ended power amplifier is shown in A. The inductor, L, and the capacitor, C, constitute a tank circuit. The inductor is tapped so that it steps up the signal voltage from the plate circuit of the driver tube. The signal is coupled from the top of the plate load to the grid through C_c. Bias is supplied to the grid through an r-f choke. The circuit of B is much the same as that of A, except that the tuned circuit is now in the grid circuit of the power amplifier, and driver plate voltage is supplied through an r-f choke. This arrangement is used when the required impedance at the grid of the power amplifier is lower than the output impedance needed in the driver stage.

(4) The circuit in C permits complete d-c isolation of the tuned circuit from the driver and amplifier stages. Capacitors $C1$ and $C2$ block the d-c voltages and at the same time couple the signal from driver plate to amplifier grid. Bias is supplied through an r-f choke. This circuit provides no means for adjusting the impedance between the grid and plate circuits.

(5) The circuit of D permits the driver stage to be neutralized. It also provides variable drive for the amplifier grid and d-c isolation for the bias and high-voltage circuits without the need for r-f chokes. The inductor, L, of the tuned circuit is split into two parts at the center, each of which is grounded separately with r-f bypass capacitors. High voltage for the driver plate therefore cannot reach the grid of the amplifier, and bias can be applied at the center tap as shown. Neutralization for the driver is obtained from the side of the tuned circuit opposite the plate through C_n.

(6) The capacitively coupled circuit in E also permits neutralization of the driver. Driver plate voltage is fed to the center tap of L through an r-f choke. A split-stator tuning capacitor, C, is used to tune the circuit to resonance. The neutralizing voltage is derived from the lower end of the coil and fed back through C_n as before; the grid drive for the amplifier is coupled through capacitor C_c from the same point. Bias for the amplifier grid is introduced through the r-f choke, as shown.

(7) In F, the tuned-plate tank of the driver is inductively coupled through a low impedance link to a tuned-grid

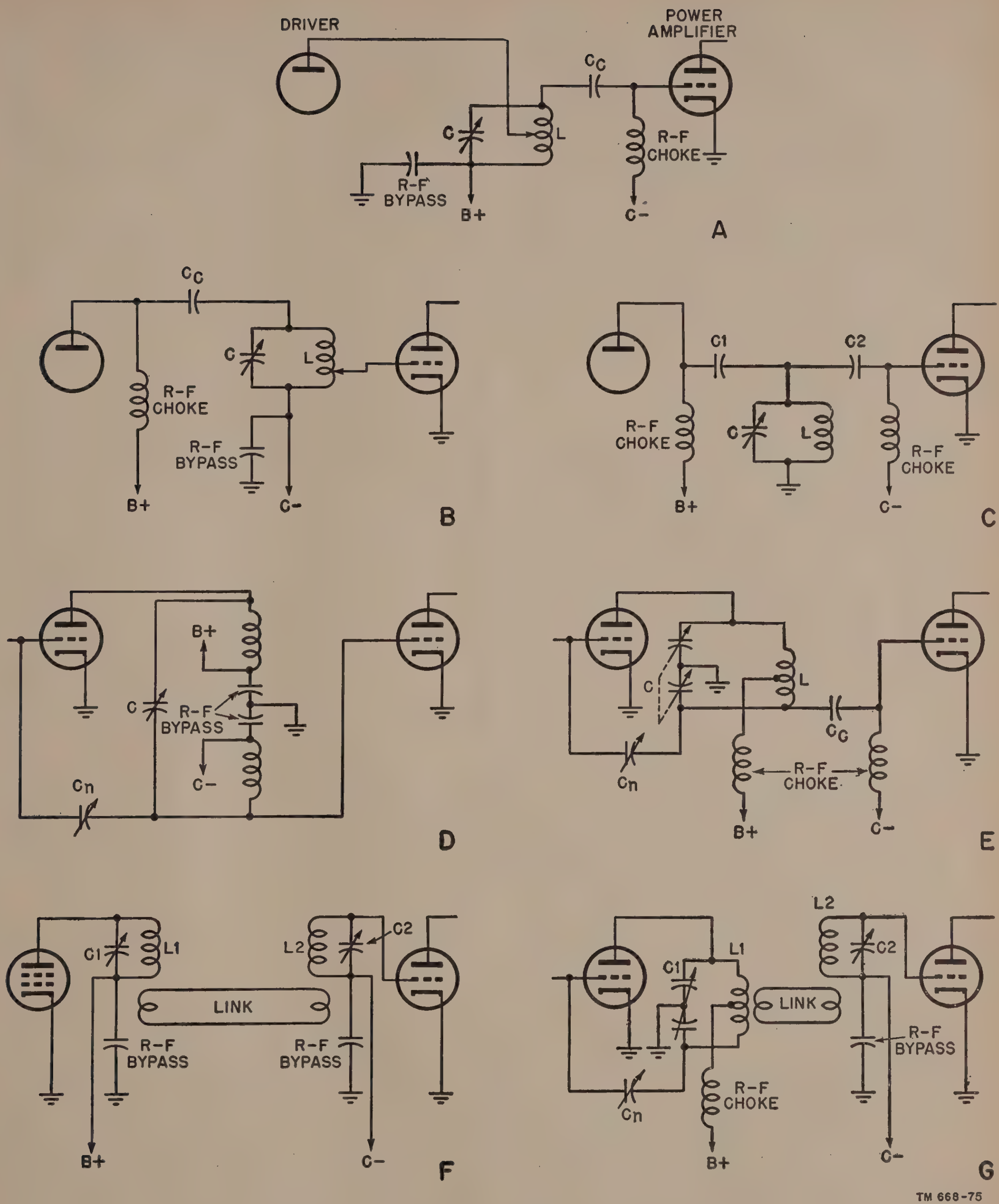


Figure 75. Practical power amplifier grid tank circuits.

circuit. The link inductance is small, and therefore the impedance of the coupling circuit is low. This minimizes losses in the transmission of driving power, and at the same time provides great flexibility in matching impedances between the driver and the power amplifier. The equivalent arrangement used with a neutralized driver is shown in G. The high voltage is applied to the tuned circuit of $L1$ and $C1$ of the driver and the neutralization is accomplished by C_n. The link circuit is slightly different in that the coupling link from the driver is positioned in the center of $L1$ rather than at its lower end.

(8) The input circuits for push-pull amplifiers (fig. 76) are variations of those used in single-ended amplifiers. The capacitive coupling arrangement in A uses a tuned-plate circuit for the driver tube formed by $L1$ and $C1$. Because each half of the split coil is out of phase with the other half, each half can supply grid drive in push-pull directly through coupling capacitors $C2$ and $C3$. A split-stator capacitor, $C1$, is used as the main tuning capacitor for the coupling arrangement. Two r-f chokes are used, one from each grid to the bias supply. Capacitor $C4$ introduces a small amount of capacitance from the lower end of the tank circuit to ground in order to compensate for the plate-to-ground capacitance of the driver tube, which appears across the upper half of the circuit. Driver plate voltage is applied through a suitable r-f choke. Neutralization of the driver stage is provided by C_n.

(9) An inductive coupling arrangement for the grid tank circuit is shown in B. The link circuit transfers energy from the driver tank, $L1$-$C1$; the out-of-phase voltages to the push-pull grids of the amplifier are developed across the split-stator capacitor, $C2$. Bias is introduced through an r-f choke. In C, the coil is center-tapped and grounded by a capacitor, and grid bias is introduced directly at the center tap. A third variation of the circuit is shown in D, where neither the split-stator capacitor nor the coil is grounded directly. $C2$ is grounded through an r-f bypass capacitor. The resistor, R, is low in value and is placed in series with the center tap of coil $L2$ and tuning capacitor $C2$. It serves to equalize any slight variation in the center tap of the coil. Bias is supplied at the center tap on $L2$.

(10) The choice of one or another of the circuits mentioned above depends on several considerations. Where the tuning capacitor is grounded directly, it must have twice the voltage rating of one that is grounded through a capacitor. In some instances, the use of an r-f choke for bias is undesirable; therefore, one of the capacitor arrangements must be used. The choice of inductive or capacitive coupling depends on the character of impedance matching and therefore indirectly on the frequency of operation. Generally, at the higher frequencies, the simpler coupling arrangements are preferred because of their lower losses. Finally, the link-coupled circuits permit the driver to be located at some distance from the amplifier and connected to it through a low-impedance transmission line.

h. Power-Amplifier Output Coupling Networks.

(1) All power amplifiers used at high frequencies must be coupled to a load circuit. Generally the impedance of the load is not the same as the plate-circuit load required by the amplifier. At the very-high frequencies where f-m is used, the amplifier is coupled to the antenna through a transmission line. In small portable units, the antenna usually is connected to the amplifier tank circuit by a different type of coupling arrangement. The impedance of the load must be matched to

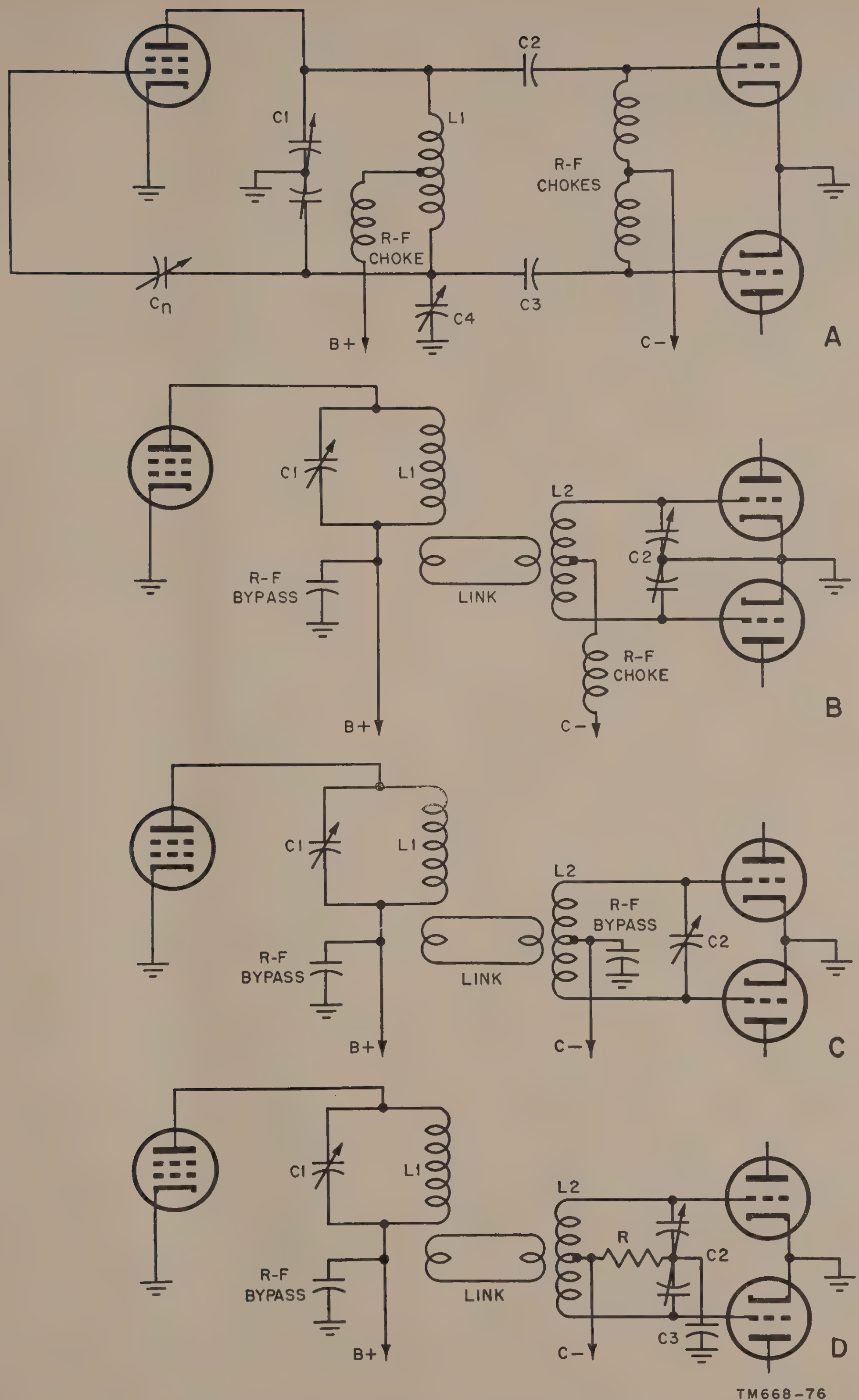


Figure 76. Grid tank circuits for push-pull amplifiers.

the plate impedance of the tube for maximum transfer of energy. There are many practical means of accomplishing this, and a variety of inductive and capacitive coupling circuits are used.

(2) The effect of the coupling network on the power amplifier is much the same regardless of the particular network involved. Analysis of the circuit shown in A of figure 77, shows that the output of the amplifier developed across the high-impedance plate tank circuit load is coupled to a low-impedance transmission line by a tuned step-down transformer. The circuit must perform the following functions:

(a) The input impedance of the coupling network must meet the tube requirements for load impedance.

(b) The transfer efficiency of the network (the ratio of output power to input power) should be as high as possible.

(c) The input impedance at all frequencies except that to be amplified should be small so that spurious frequencies generated in the amplifier are minimized.

(d) The selectivity of the network should suppress unwanted frequencies and prevent their transfer to the load. The selectivity must not be so high that outer side bands of the frequency-modulated signal are reduced in amplitude.

(3) Capacitor $C2$ in A is adjusted so that it resonates with $L2$ at the center operating frequency. The coupling between $L1$ and $L2$ then is adjusted to present approximately the desired load impedance to the tube. Finally, $C1$ is varied until the circuit presents a purely resistive impedance to the tube. At this point, the d-c plate current of the amplifier is a minimum because the power factor is unity. The tuning curve of d-c plate current versus capacitance is shown in B.

(4) The load impedance presented to an amplifier by a parallel circuit at resonance is equal to the product of the operating Q and the inductive reactance. The actual Q of the inductance is many times higher than the operating Q since the load that is reflected into the parallel circuit by transformer action reduces the effective Q of the circuit. This is equivalent to placing additional resistance R_{eq} in series with the actual resistance, $R1$, of coil $L1$, as shown in the equivalent circuit of C. The effect of coupling a load to a parallel-resonant circuit is to decrease the effective Q. For high efficiency,

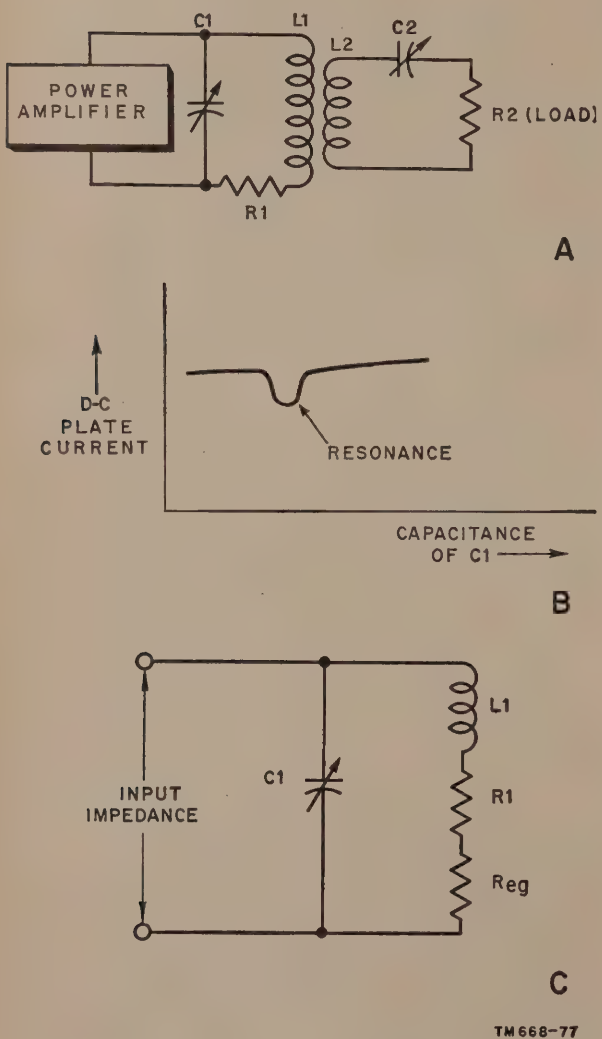


Figure 77. Ideal power amplifier output coupling network.

the unloaded Q of the tank circuit should be as high as possible. The loaded Q then should be made as low as possible. However, this conflicts with the requirements of selectivity and low impedance to spurious frequencies. Therefore, a compromise value is chosen. Typical values involve unloaded Q's of 200 and loaded Q's of 10 to 12 in v-h-f f-m transmitters. The transmitter efficiency of the tank circuit can be shown to be as follows:

$$\text{Transfer efficiency} = \frac{Q_o - Q_L}{Q_o} \times 100 \text{ percent}$$

where Q_o is the unloaded Q of the parallel-resonant circuit and Q_L is the loaded Q. For the values given above, the efficiency is

$$\frac{200-12}{200} \times 100 = 94 \text{ percent}$$

i. Practical Transmitter Inductively Coupled Tank Circuits.

(1) A large variety of practical circuits have been devised which present the proper load impedance to the power amplifier when connected to the transmission line or antenna. Some of the coupling circuits that have not been discussed are illustrated in figure 78. The simple, parallel-resonant tuned circuit in A frequently is used for single-ended tetrode amplifiers. The circuit is shunt-fed, the plate voltage being applied in parallel with the tank circuit. The d-c plate voltage applied through the inductance is fed through an r-f choke which effectively isolates the power supply. The tank circuit, $C1$-$L1$, is coupled to the plate by capacitor $C2$. The advantage of this circuit lies in the removal of all d-c voltages from the tuning capacitor. This means a lower value of total voltage across this component, with correspondingly smaller size and lighter weight. A major shock hazard from contact with an exposed portion of the tank circuit is removed. However, the possibility of a bad r-f burn always exists.

(2) Although the circuit in A of figure 78 is satisfactory when used with tet-

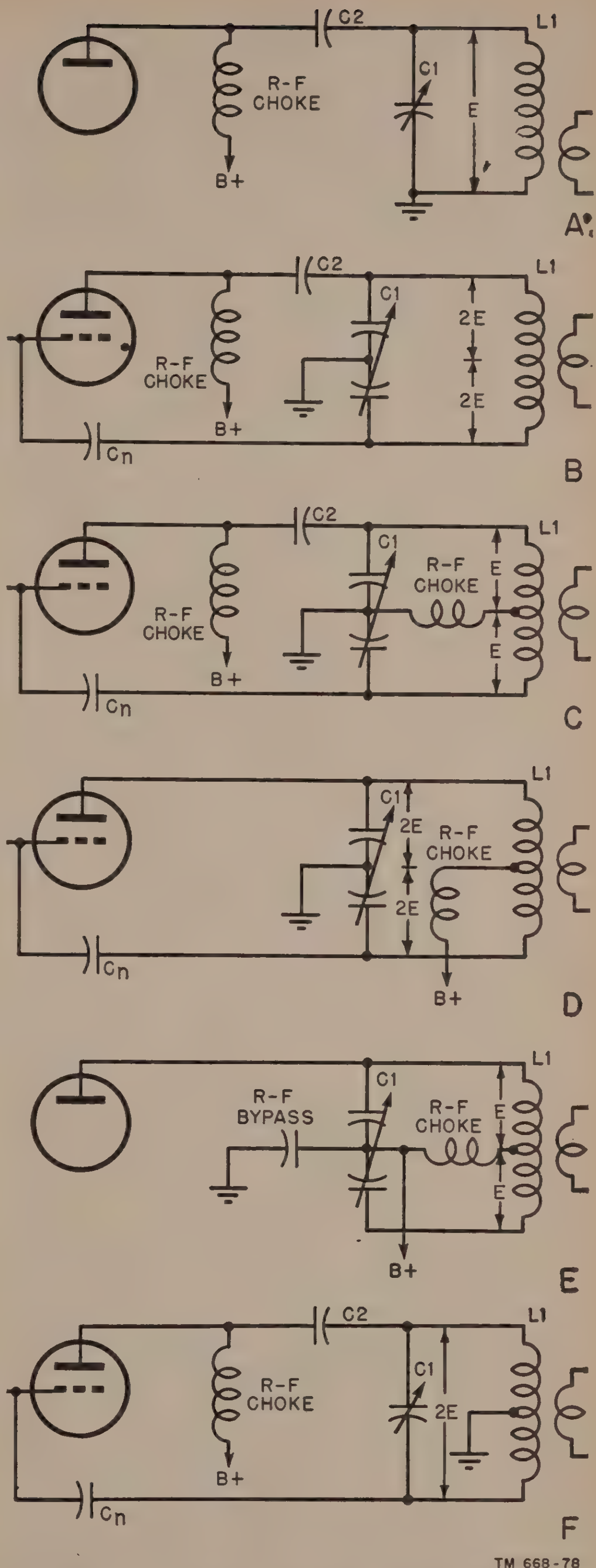


Figure 78. Practical plate-tank circuits for class-C amplifiers.

rodes and grounded-grid triodes, it provides no means of neutralizing an ordinary grounded-cathode triode. In B, the capacitor C, has a split stator with the rotor directly grounded. This permits an out-of-phase voltage to be taken from the lower end of the coil and returned to the input through the neutralizing capacitor, C_n. Plate voltage is applied as before through an r-f choke. The blocking capacitor, $C2$, prevents the plate voltage from reaching the tuned circuit. However, the split-stator capacitor effectively divides the circuit in two parts, and an r-f peak of twice the d-c plate voltage can appear across each. This requires a physically large capacitor and limits the use of this circuit.

(3) The problem of excessive capacitor voltage is solved in the circuit in C, where shunt feed is retained along with the neutralizing circuit. However, the coil is center-tapped and a small r-f choke places the center of the coil at the d-c ground potential. The over-all voltage across each half of the circuit becomes that of the a-c voltage alone, and permits the size of the tuning capacitor plate spacing to be reduced.

(4) Another tank circuit which allows the neutralization of a single-ended stage is shown in D. This circuit is series-fed through an r-f choke to the center of the tapped tank coil. A split-stator capacitor is used for tuning and the neutralizing voltage is taken from the lower end of the coil through C_n. This circuit has the same faults as the one in B, since the r-f equivalent of twice the plate voltage appears across each half of the circuit. It can be remedied by the circuit of E, where the rotor of the tuning capacitor is left ungrounded for d-c, but is bypassed for r-f by a capacitor. The power-supply voltage is applied to the rotor to equalize the voltages that would otherwise be built up across the tank. This results in a severe operating hazard unless the shaft of the tuning capacitor is well insulated from the tuning knob.

(5) The circuit at F is a modification of this arrangement, permitting the use of shunt feed and doing away with the necessity for a split-stator capacitor. Plate voltage is applied through an r-f choke, and the d-c voltage is prevented from reaching the tank by a blocking capacitor, $C2$. The tank inductance itself is grounded directly at the center tap, making this circuit desirable when a grounded coil is needed. This is the case where the coils are selected by a rotary indexing switch when changing frequency of operation over a wide range.

(6) Almost all of the preceding circuits for single-ended stages have their push pull counterparts. Some of these are shown in figure 79. The circuit in A is the push-pull counterpart of the simple resonant tank. A split-stator capacitor is used with the push-pull version, and the rotor is grounded for r-f through a bypass capacitor. Plate voltage is series-fed to the center of the inductor through an r-f choke. To reduce the voltage across each half of the tuning capacitor, the plate voltage sometimes is connected to the rotor. The shock hazard introduced by this can be avoided by grounding the rotor of the tuning capacitor directly, and applying the plate voltage through an r-f choke, as in B. This circuit has an r-f peak of twice the d-c plate voltage appearing across each section of the capacitor.

(7) A third alternative, which is less desirable than the other two, is shown in C. Here, plate voltage is applied at the center of the inductor, which is grounded by a bypass capacitor at that point. A single-section tuning capacitor is used, insulated entirely from d-c or a-c ground. An r-f peak of twice the d-c plate voltage appears across the tank. Many other variations are possible following the general principles used in each example.

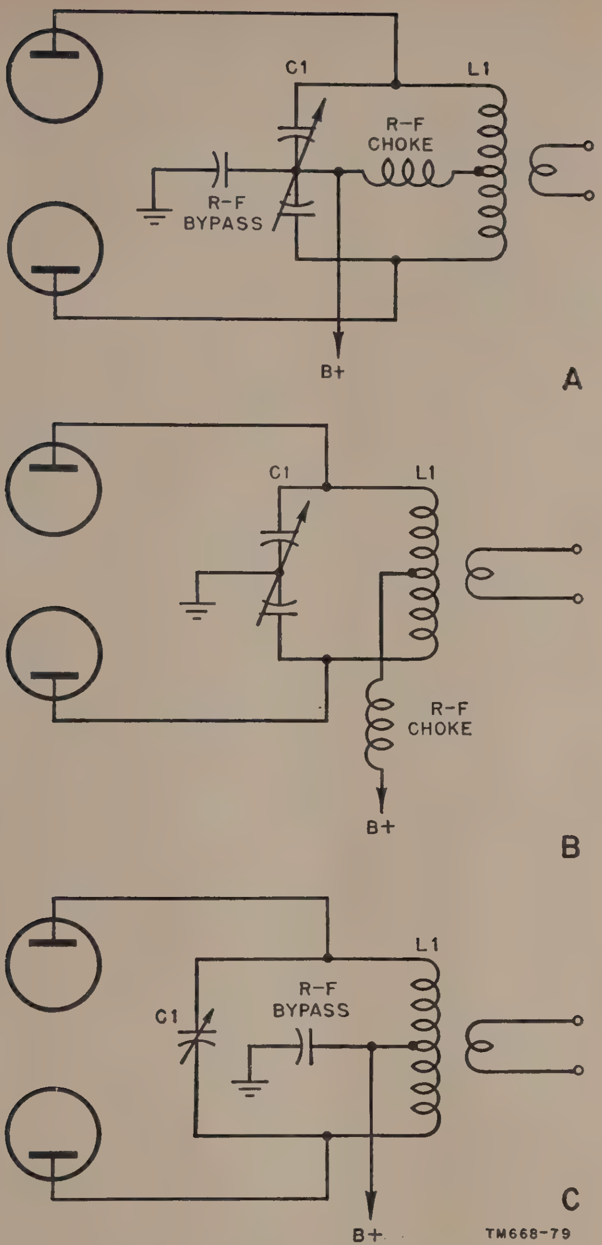


Figure 79. Push-pull plate tank circuits.

j. Antenna Matching Tank Circuits.

(1) In small, portable transmitters, it is common to find the antenna connected directly to the tank circuit of the transmitter with no intervening transmission line. Because the impedance of this kind of antenna can vary over a considerable range, it must be matched directly to the power amplifier by means of a tank circuit that can compensate for a wide range of impedance. Three basic circuits, adaptable to single-ended and push-pull tank circuits alike, which provide this variable impedance matching are shown in figure 80.

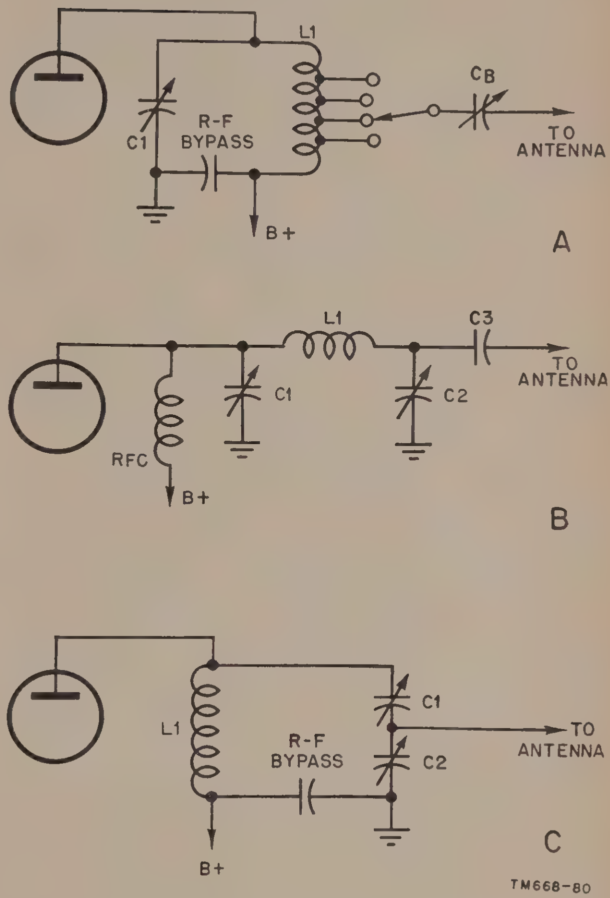


Figure 80. Antenna-matching tank circuits for power amplifiers.

(2) A series-fed, parallel-resonant, single-ended tank with grounded tuning-capacitor rotor and bypassed inductor is illustrated in A. Instead of coupling the antenna inductively to the tank, it is tapped directly to the coil through a blocking capacitor. Since the lower end of the tuned circuit is grounded effectively, the impedance to ground at that point must be zero. As the tap is moved up on the coil, the impedance rises until it reaches the ultimate value of the tank circuit impedance. In a

practical transmitter, the coil is tapped at intervals and a rotary switch is used to select the tap which gives the proper value of coupling. Because the d-c plate current in the amplifier stage increases with increased loading, the tap can be set at the point which gives the required d-c current in the stage when the tuned circuit is resonated. If the blocking capacitor, C_B, between the tap and the antenna is made variable, a further adjustment between separate taps can be obtained.

(3) One of the most frequently encountered variable matching networks for the output of an r-f amplifier is the circuit in B. The plate voltage fed through the r-f choke is prevented from reaching the antenna by blocking capacitor $C3$. The simple pi-network of $C1$, $L1$, and $C2$ is capable of matching a wide range of impedances, and operates as a voltage divider. The combination of $L1$ and $C2$ forms the divider circuit which develops higher or lower voltages at the output terminal. $C1$ then tunes the combination of $C2$ and $L1$ to resonance at the operating frequency. Depending on the relative values of $C1$ and $C2$, a voltage much lower than the a-c plate voltage can be developed. Consequently, this circuit can match an extremely wide range of impedances. In addition to matching purely resistive loads, the circuit also can compensate for a certain amount of reactance. This is important when using short antennas which introduce considerable capacitive reactance.

(4) A variation of the pi-network, in which one of the capacitors is not grounded, is shown in C. Capacitors $C1$ and $C2$ themselves form the impedance-matching voltage divider. The circuit cannot match as wide a range of impedances as the pi-network can, and it is further limited because the rotor of $C1$ must be carefully insulated. The response of these circuits to harmonics of the fundamental frequency is poor, which is a desirable feature. The pi-network does not discriminate against signals below operating frequencies. This makes it undesirable to use if the amplifier is driven directly by a frequency multiplier.

k. Parasitic Oscillation, Adjustment, and Neutralization.

(1) Many different types of input and output circuits for class C f-m power amplifiers have been described. At first glance, the large variety of choices available makes it seem difficult to understand why a particular circuit is chosen. Not all combinations of input and output circuits can be used together successfully, since some of them permit the amplifier stage to oscillate at frequencies that are relatively unrelated to the frequency to which it is tuned. These *parasitic oscillations* are distinct from the sort of oscillation that occurs in an amplifier which is improperly neutralized or one in which the input circuit is not shielded sufficiently from the output. They are undesirable because they cause the transmission of spurious signals, thus impairing the efficiency of the amplifier.

(2) The most noticeable features of parasitic oscillation in an amplifier are erratic tuning and the radiation of spurious frequencies. When an amplifier is operating properly, the d-c plate current dips sharply as the tank circuit is tuned through resonance. This plate current minimum also corresponds to maximum power output. If a tetrode is operating normally, the plate-current change may not be too great, but the screen-current dip will be significant. With parasitic oscillation, the plate current may not dip at all; the minimum may not correspond to maximum power output; or several dips may appear in the tuning range. Since the symptoms presented by a stage which is not properly neutral-

ized are somewhat similar, it is difficult to tell the two effects apart unless neutralization is checked first.

(3) All parasitics are attributable to the development of resonant circuits in connection with the tube elements in such a way as to permit enough feedback to sustain oscillation. They may occur at either high or low frequency. Parasitic oscillations occurring at much lower than operating frequencies usually are caused by the resonant condition of an r-f choke in the circuit, since the r-f chokes are the only inductors with sufficient inductance to resonate with various circuit capacitances at low frequencies. High-frequency parasitics can be traced to a much wider variety of causes. Among these are spurious high-frequency resonant conditions in tank-circuit inductances; resonant circuits built up in lead inductances and stray, or tube, capacitances, and resonant conditions built up in bypass and blocking capacitors. Moreover, the parasitic circuit need not involve the final amplifier alone. The driver stage is frequently an important part of the parasitic feedback circuit which permits oscillation.

(4) A recurrent type of high-frequency parasitic oscillation is caused by a form of tuned-plate, tuned-grid oscillator in a simple single-ended amplifier like that of figure 81. The parasitic path is shown in heavy lines. At relatively high frequencies, the tank-circuit inductance acts like an r-f choke, and the capacitors and the leads from them form the equivalent of parallel-resonant circuits. The shielding effect of the screen grid in a tetrode is not sufficient at extremely high frequencies. Therefore, energy can feed back to the grid circuit from the plate at high frequencies if both of the parasitic resonant circuits are almost the same in frequency. The difficulty can be cured by inserting a parallel inductance and resistance in the grid or plate lead. This detunes one of the parasitic circuits sufficiently to prevent oscillation. Another method is to insert a small resistance in series with circuit leads to introduce sufficient loss to stop oscillation. A third alternative is to incorporate a tuned parallel-resonant trap that actually inserts a very high impedance in the parasitic frequency path. In addition to the trap circuit, it is common to find small high-frequency capacitors connected from plate and control grid to cathode. These capacitors effectively bypass the harmonic path.

(5) Certain circuit combinations have been found to be troublesome. For example, r-f chokes rarely are used in both the grid and the plate circuit of a triode, since they cause a low-frequency tuned-plate, tuned-grid oscillation (B of fig. 81). For this reason, shunt-fed circuits are avoided whenever possible, since they encourage parasitic difficulties. In high-gain screen-grid amplifiers, the selection of the screen bypass capacitor becomes very important. The substitution of a different type when servicing a unit often leads to serious instability. Similarly, the choice of cathode or filament bypass capacitors is also a more critical matter than the circuit diagram tends to indicate. When replacing any of these components in a transmitter, use the *exact duplicate* of the discarded component, and *pay careful attention* to lead dress and parts placement.

(6) In push-pull circuits, there are many more possibilities for the development of parasitic oscillation. However, the oscillation usually can be traced to one of the standard types of oscillators operating in conjunction with the tube leads or r-f chokes. Remedies similar to those found in single-ended stages are used. An attempt always is made to space the spurious resonant circuits in such a way that none of them occur at the same frequency in the output and input circuits. As such,

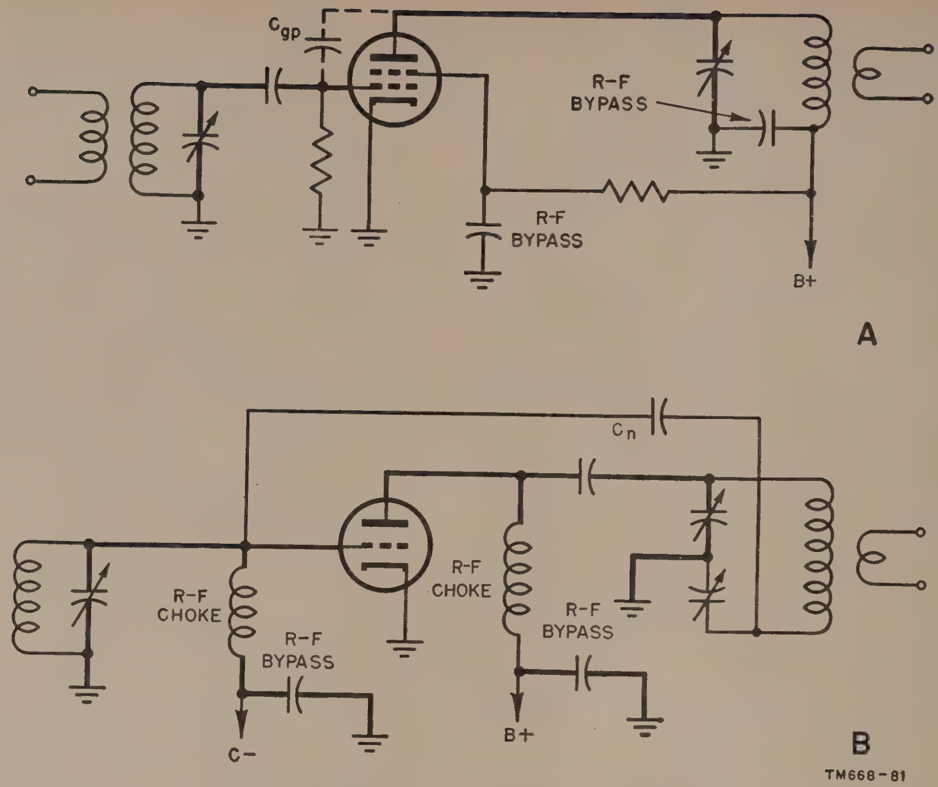


Figure 81. Parasitic oscillation circuits.

the circuit diagram of a power amplifier does not always tell the whole story of its operation. Many features are involved in the stability of the circuit that do not appear in the formal schematic, since they have to do with lead length, choice of wire size, placement of parts, and type of component.

(7) In f-m transmitters, improper neutralization can lead to a number of difficulties. The amount of frequency deviation can be changed by a final power amplifier on the verge of oscillation. In addition, such amplifiers are difficult to adjust for optimum power output and performance. Since the f-m signal is steady in amplitude, there should be no fluctuation of any of the d-c voltages or currents in the final amplifier when the transmitter is modulated. Any variation in rectified grid current indicates either overmodulation or instability in the driver stages. Fluctuation in d-c plate current points to parasitics or improper neutralization. Only a stable, properly tuned amplifier is capable of providing satisfactory performance.

Section II. AUTOMATIC-FREQUENCY CONTROL

42. Frequency Control

a. Description and Purpose. In military transmitters and receivers precise maintenance of assigned frequencies is imperative. Some transmitters must be capable of continuous frequency variation over an assigned band of frequencies and often are combined with receivers in the same unit. Since these receivers must be kept on exactly the same frequency as the transmitter, a control system must be used to compensate for any variation in frequency caused by vibration, temperature change, or humidity.

The control system is essentially an error-correcting system and usually is referred to as automatic-frequency control, or afc.

b. General Afc Systems. In an afc system, some of the output voltage is sampled and compared with a constant-frequency source, and any frequency difference that exists results in automatic correction of the frequency of the master oscillator. The functional arrangement of all afc systems is pictured in the block diagram of figure 82.

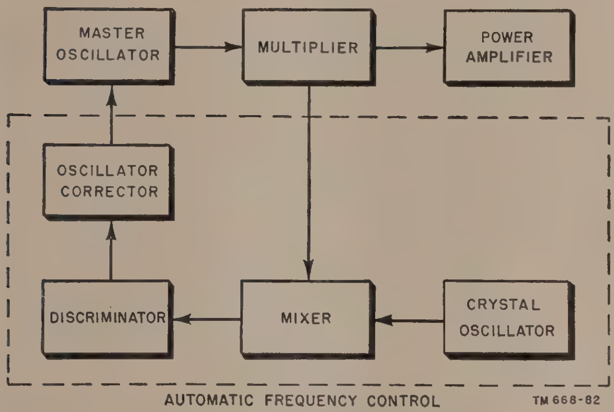


Figure 82. Functional block diagram of afc system.

(1) The classification of different types of afc depends on the means used to control the frequency of the oscillator. Oscillators which are frequency-modulated by a reactance or Miller effect respond to d-c voltages, and produce corresponding frequency changes. The correction voltage that is applied to the oscillator must be a d-c voltage directly proportional to the frequency difference between the output of the master oscillator and a standard crystal oscillator. The comparator, therefore, must produce a d-c voltage that corresponds to the difference frequency. Such a device is called a *discriminator*.

(2) Control of the frequency of an oscillator through the application of direct voltage to the modulator has certain disadvantages from the standpoint of precision. D-c amplifiers are unstable in respect to changes in tubes and line voltages. Moreover, if the control system fails, the frequency of the oscillator changes drastically because of the abrupt d-c voltage change at the modulator. From the standpoint of reliability this is undesirable. However, the d-c control systems are simply constructed and, despite their inherent disadvantages, they are widely used.

(3) To overcome the disadvantages of the d-c control system, a mechanical element can be inserted in the system to act as the oscillator frequency controller. This is usually a two-phase motor, whose rotor position depends directly on the phase relation of the voltages across its field windings. The motor has a shaft with a variable capacitor or inductor attached. The capacitor or inductor is connected in the circuit so that the reactive element varies the frequency of the master oscillator. This method of afc usually is applicable only to large fixed stations where weight and size are not important. For extremely high frequencies, where d-c control systems do not provide the required accuracy, the use of a motor-positioning system is frequently the only alternative. The motor positioning system has the advantage, in the event of failure of the control system, that the motor shaft does not turn. Therefore, the correction reactance produced does not change, and the frequency is not disturbed. Of course, if the failure lasts over a considerable period of time, the oscillator itself shifts frequency because of temperature changes and other causes.

(4) Motor control systems and d-c systems have varying speeds of response to errors of frequency. Because of the mechanical inertia of a motor, motor systems generally cannot respond quickly to changes in frequency such as the change caused by modulation of the carrier. However, it can respond much more accurately to slow variations. The d-c systems are pre-

vented from responding to modulation by use of circuits with long time constants.

(5) Receiver interlock systems, used in small portable units with single-dial control, are generally of the d-c type. Interlock systems use the same master oscillator for the receiver and the transmitter. Mixer circuits using crystal-controlled oscillators produce the required frequencies for the local oscillator of the superheterodyne receivers and for the actual master control of the transmitters. This system can be further refined by using the receiver to keep the transmitter locked to the frequency of another station.

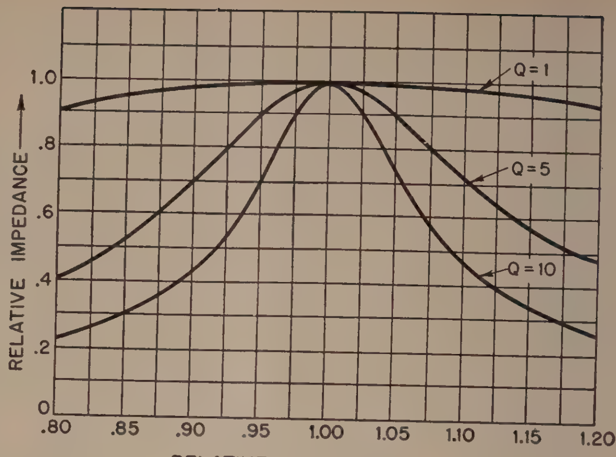


Figure 83. Response curves of tuned resonant circuits.

43. Discriminator

a. General.

(1) The *discriminator* is a device for producing a d-c voltage which is proportional to the frequency of an input signal. The polarity of the voltage produced depends on whether the frequency is higher or lower than the frequency to which the discriminator is tuned. The response curve of a tuned resonant circuit (fig. 83) shows that the sides of the curve approach straight lines as the Q of the coil is increased. If a signal of variable frequency is coupled to a tuned circuit, the voltage produced across it will depend on the relation of the frequency of the coupled voltage to that of the tuned circuit. The voltage also depends on the Q, since Q defines the sharpness of resonance so that frequencies farther away from resonance produce less voltage across the circuit.

(2) If the a-c voltage that exists across a resonant circuit is rectified by a diode, a d-c voltage proportional to the amplitude of the a-c voltage is produced. If the amplitude of the a-c voltage varies with the displacement of the applied frequency from the actual resonant frequency of the tuned circuit, the d-c voltage will increase or decrease.

b. Double-Tuned Discriminator.

(1) Figure 84 shows a double-tuned discriminator consisting of tuned circuits $T1$, $T2$, and $T3$, diode rectifiers $D1$ and $D2$, and the filter networks, $R1C1$ and $R2C2$. The secondaries, $T2$ and $T3$, are tuned to resonate at different frequencies; one is tuned above the carrier frequency and the other an equal distance below the carrier frequency. This provides equal voltages at the center frequency, as shown in the response curve of figure 85. When an r-f voltage that is constant in amplitude and varying in frequency is applied to $T1$, the voltages induced in $T2$ and $T3$ will be 180° out of phase, and alternate voltage polarities will appear at the plates of $D1$ and $D2$. These induced voltages increase and decrease in amplitude with the changing frequency. For example, assume that $T2$ is tuned to a frequency higher than the center frequency. As the induced voltage approaches the resonant frequency of $T2$, its amplitude increases in a positive direction. If $T3$ is tuned to a lower frequency than the center frequency, as the induced voltage approaches the resonant frequency of $T3$ its amplitude increases in a negative direction. When the induced voltage goes positive at the plate of $D1$, current flows in the circuit $D1$, $T2$,

and $R1$ and a voltage proportionate to the change of frequency appears across $R1$. As the induced voltage goes positive at the plate of $D2$, current flows in circuits $D2$, $T3$, and $R2$, and a voltage proportionate to the change in frequency appears across $R2$. Capacitors $C1$ and $C2$ across the rectifiers filter out any a-c variations, and permit a d-c voltage to be built up across the load resistors.

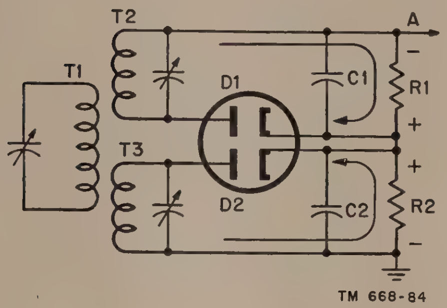


Figure 84. Double-tuned discriminator circuit.

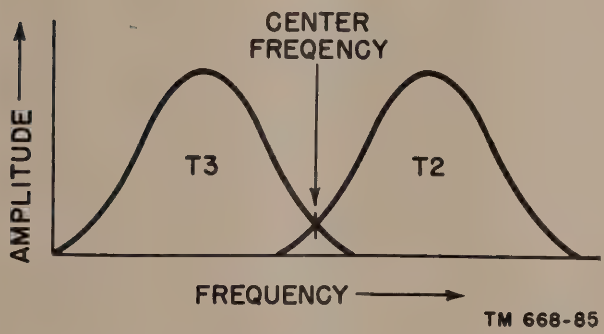


Figure 85. Response curve of double-tuned discriminator.

(2) With the connection of the rectifier as shown, the voltages across the individual load resistors oppose one another because the cathodes are both at the same potential. Therefore, the total voltage between the top of $R1$ and ground depends on the relative value of the voltages across $R1$ and $R2$. Since the voltage across the individual load resistors depends only on the frequency of the applied signal, when the frequency is higher than the center frequency, the voltage developed by the diode connected to the circuit tuned above the center frequency is higher than that developed across the other tuned circuit. Similarly, if the applied frequency is lower than the center frequency, the diode connected to the low-frequency tuned circuit produces the larger voltage across its load resistor. If the frequency of the applied signal is exactly at the center frequency, the voltage across the load resistors is equal, and the total output voltage is zero.

(3) When the applied frequency is higher than the center frequency, more voltage is developed across $T2$, a greater d-c voltage appears across $R1$, and A becomes more negative. With the carrier at the center frequency, the voltage at A, in respect to ground, becomes zero. When the frequency swings lower than the center the voltage developed across $T3$ is greater than that produced across $T2$, a greater d-c volt-

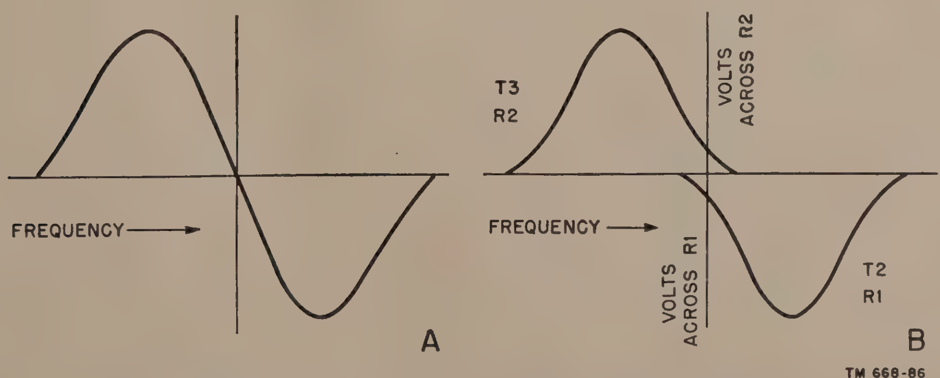


Figure 86. Output voltage of double-tuned discriminator.

age appears across $R2$, and point A becomes positive in respect to ground. Therefore, as the applied signal swings from below to above the center frequency, the voltage at A goes from positive to zero to negative. This results in the curve of output voltage versus frequency shown in A of figure 86. The voltage across the individual load resistors in respect to frequency is shown in B.

c. Use of Double-Tuned Discriminator.

(1) The potential developed at the output of a double-tuned discriminator can be used as a frequency-correction voltage by applying it to the grid of a reactance-modulator tube. When the discriminator voltage changes polarity, the transconductance of the tube is increased or decreased and the frequency of the oscillator shifts. For example, assume that the reactance modulator is connected to inject capacitance into the oscillator circuit. When the transconductance is reduced, less capacitance is injected and the oscillator frequency increases a small amount. As the frequency applied to the discriminator changes and the voltage output becomes positive, the transconductance of the reactance modulator is increased. This injects more capacitance across the oscillator tank circuit and the frequency of oscillation is lowered.

(2) If the discriminator is tuned so that an increase in frequency of the oscillator produces a positive voltage, the discriminator voltage applied to the reactance-modulator tube tends to return the oscillator to the center frequency. Similarly, a decrease in frequency will cause the oscillator to return to its normal frequency. Since all of the operations take place at the oscillator frequency, and since the frequency at which the system becomes stable is the center frequency of the discriminator characteristic, this is a crude control system. The inductance and capacitance in resonance are no more likely to be stable than the inductance and capacitance of the oscillator tank circuit itself. However, this is the only method of afc that can be used in some ultrahigh-frequency systems. Because the over-all accuracy is totally dependent on the center of the discriminator frequency characteristic, the discriminator must be more stable than the oscillator for accurate frequency control.

(3) A high-accuracy afc system should keep the center frequency of the master oscillator as stable as a crystal oscillator. This can be accomplished by comparing the frequency of the applied signal with the frequency of a crystal. The frequency difference between the crystal and the master oscillator, as produced in a mixer, depends on the absolute values of both frequencies. Therefore, if the crystal oscillator is assumed to be absolutely stable, the difference frequency depends only on that of the transmitter frequency. This difference frequency is applied to a discriminator operating at a much lower frequency than the oscillator and the transmitter. If the frequency at which the discriminator operates is made low enough, the result of a variation in its tuned circuits is only a few kilocycles. Therefore, if the crystal frequency, or any multiple of it, is mixed with the master-oscillator ouput to produce a low difference frequency, the low-frequency discriminator tuned circuits will cause only a small error, whereas the over-all variation of difference frequency caused by drift in the oscillator will be much larger. The departure of the transmitter frequency from the difference frequency produced by the mixer determines the correction voltage. The output of the discriminator feeds the reactance modulator tube, which brings the master oscillator back to the center frequency. The over-all stability is nearly that of the crystal oscillator itself, differing only by the de-

parture of the discriminator itself from the low-center frequency. This is a departure of approximately a few hundred cycles per second.

(4) The frequency of the master oscillator shifts with the applied audio signal, but the discriminator is so constructed that it responds only to changes in frequency that are much slower than the audio variations. The voltage at the output of the discriminator varies at a rate equal to the rate of change of frequency of the master oscillator—that is, at the audio rate. It also varies at a much slower rate because of oscillator drift. By placing a low-pass filter after the discriminator, only the slow drift variations can reach the reactance tube. This filter cuts off all voltage changes with a rate that is equal to or higher than the audio frequencies used.

d. Phase Discriminator.

(1) Figure 87 shows the schematic diagram of another type of discriminator circuit used in afc circuits. This is known as the phase discriminator. The voltage output of the discriminator depends on the phase relations in the circuit. Transformer *T*1 couples the input from the crystal oscillator to the diode plates producing equal and opposite voltage *E*2 and *E*3 on these plates. The phase relationship of these two voltages is shown in A of figure 88. The crystal oscillator and transmitter frequencies are reduced so that these voltages are generally of a frequency just above the audio range. Since the input to *T*1 is crystal-controlled, it can be considered as stable. Therefore, the frequency and relative phase of *E*2 and *E*3 never change. The input from the master oscillator of the transmitter is injected into the primary of *T*2. This produces a voltage, *E*1, across its secondary, which is exactly 90° out of phase with *E*2 and *E*3 when the frequency of the input from the crystal is the same as the frequency from the master oscillator. The upper diode receives a voltage equal to the sum of *E*1 and E2; the lower diode receives voltage equal to the sum of *E*1 and *E*3. The diodes rectify the signals and d-c voltages appear across the load resistors. When the signal voltages are equal, the output voltages across the resistors are equal, and the total voltage across both resistors in respect to ground is canceled out.

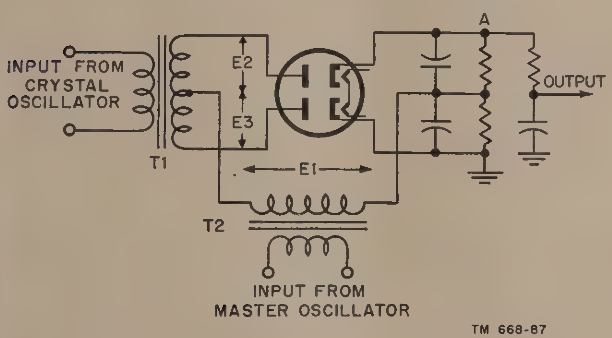


Figure 87. Basic phase discriminator.

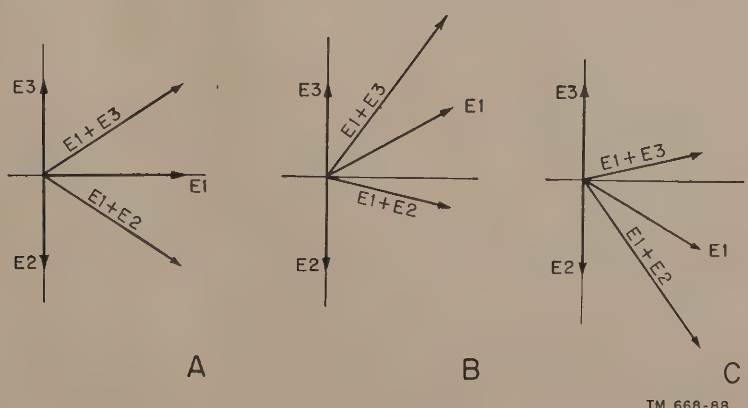


Figure 88. Vector relations in phase discriminator.

(2) If the frequency of the master oscillator increases, the vector relationships change as shown in B. The increased frequency is equivalent to a phase shift of $E1$ in respect to $E2$ and $E3$. The resultant vectors, $E1$ plus $E3$ and $E1$ plus $E2$, therefore change in length, as shown, and the ratio of the voltage across the diodes changes. The voltages across the load resistors (fig. 87) are now unequal and a voltage that can be fed back into the modulator to correct the drift of the master oscillator is produced at point A.

(3) Similarly, if the frequency of the master oscillator decreases, the phase of $E1$ changes, as shown in C of figure 88 and a voltage of the opposite polarity is produced across the diodes. The voltage across the lower diode is now less; therefore, a less positive voltage is produced at point A (fig. 87). The result is a change in the polarity of the over-all voltage to that produced when the frequency of the master oscillator shifts upward in frequency. The correction voltage applied to the proper circuits tends to return the system to a condition where $E1$ is 90° out of phase in respect to $E2$ and $E3$. Consequently, the frequency of the master oscillator is dependent on that of the crystal oscillator. Since the discriminator depends only on phase relationships and not on the absolute frequency of the secondaries of the transformers, the tuning of the discriminator has less effect on the stability of the system. However, divider circuits are needed to reduce the transmitter and crystal oscillator frequencies and they are considerably more complex than the simple mixer system.

e. Modified Phase Discriminator.

(1) The complex divider circuits necessary have led to a modification in the basic phase discriminator. Essentially, this modification consists in obtaining the voltage that corresponds to that produced by $T2$ in figure 87 from the same circuit to which the split secondary of $T1$ is connected. This modification is shown in figure 89. A tuned circuit is used for the split transformer, the signal voltages across the two halves of the secondary are still $E2$ and $E3$, and the diode load circuits and their operation are the same as before. The voltage present across the primary winding is also present across L and the voltage, $E1$, is obtained from across this inductor. Therefore, the secondary system receives its voltages in two ways—by inductive coupling, and through coupling capacitor C. The voltage is the same as the voltage, $E1$, across the primary and is 180° out of phase with the total voltage, $E2$ plus $E3$, induced across the secondary circuit. The voltage drops across each half of the split secondary are, in turn, 90° out of phase with the applied voltage. However, they are 180° out of phase with each other, since the circuit is a tapped transformer, and therefore, the voltages $E2$ and $E3$ are developed 90° out of phase with the primary voltage.

(2) When the frequency of the carrier departs from the center of the tuned-circuit resonance curve, the phase relationships across the tuned circuit change, and the over-all d-c output changes. Consequently, the d-c output of the discriminator varies with the applied frequency. However, operation is dependent on the resonant frequency of the discriminator secondary, although this discriminator is much easier to tune and adjust than the double-tuned discriminator, because only one resonant circuit is used.

f. Pulse Discriminator.

(1) Another type of discriminator that can be used in afc systems—the *pulse discriminator* — uses pulses rather than continuous sine-wave signals. The pulse discriminator distinguishes between two pulse signals of different repetition rates, and produces a volt-

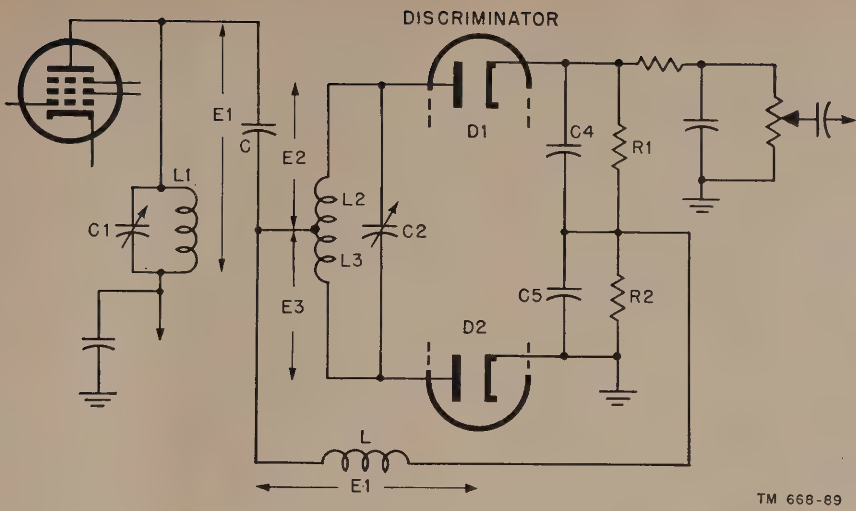


Figure 89. Modified phase discriminator.

age pulse proportional to their difference. It is necessary to convert the frequency of the master oscillator to a series of pulses with a repetition rate proportional to frequency. The repetition frequencies are compared in the pulse discriminator and an output is produced if the repetition rates differ. This output is converted to a d-c voltage and applied to the modulator, thereby shifting the frequency and bringing the two repetition rates back into synchronism.

(2) The accuracy of the pulse system is much greater, in terms of timing, than that of a system using sine waves. A pulse begins at a precise instant, whereas a sine wave does not, because the rate of voltage change at the beginning of a pulse is much greater than the rate of voltage change at the beginning of a sine wave. A pulse discriminator responds to smaller changes in the repetition rate than an equivalent phase discriminator, which responds to changes in the frequency of the input signals. This system of stabilization has the advantage of requiring no frequency-divider circuits. Moreover, no tuned circuits of any kind are used in the actual stabilization loop. This means that there need be no tuning or operating adjustment of the stabilization circuits.

(3) The over-all block diagram of the stabilization circuit is shown in figure 90. Some of the f-m signal from one of the multiplier stages in the transmitter is tapped off and fed into a buffer amplifier. This amplifier applies the signal to the input grids of two mixer tubes. A crystal oscillator, operating at the same frequency as the input signal, is applied also to the mixer grids. However, the signal from the oscillator is passed through two R-C networks, each of which shifts the phase by 45°. To accomplish this, the value of the resistance in each R-C network is made equal to the reactance of the capacitor at the crystal frequency. The upper network shifts the phase forward by 45° and the lower retards it by the same amount. Therefore, the dual output of the networks differs in phase by 90°.

(4) The mixer plate circuits select the frequency which is the difference between the signal from the buffer amplifier and the signal from the oscillator. When no modulation is applied to the transmitter and the frequency of the master oscillator is stable, the frequency from the buffer oscillator and

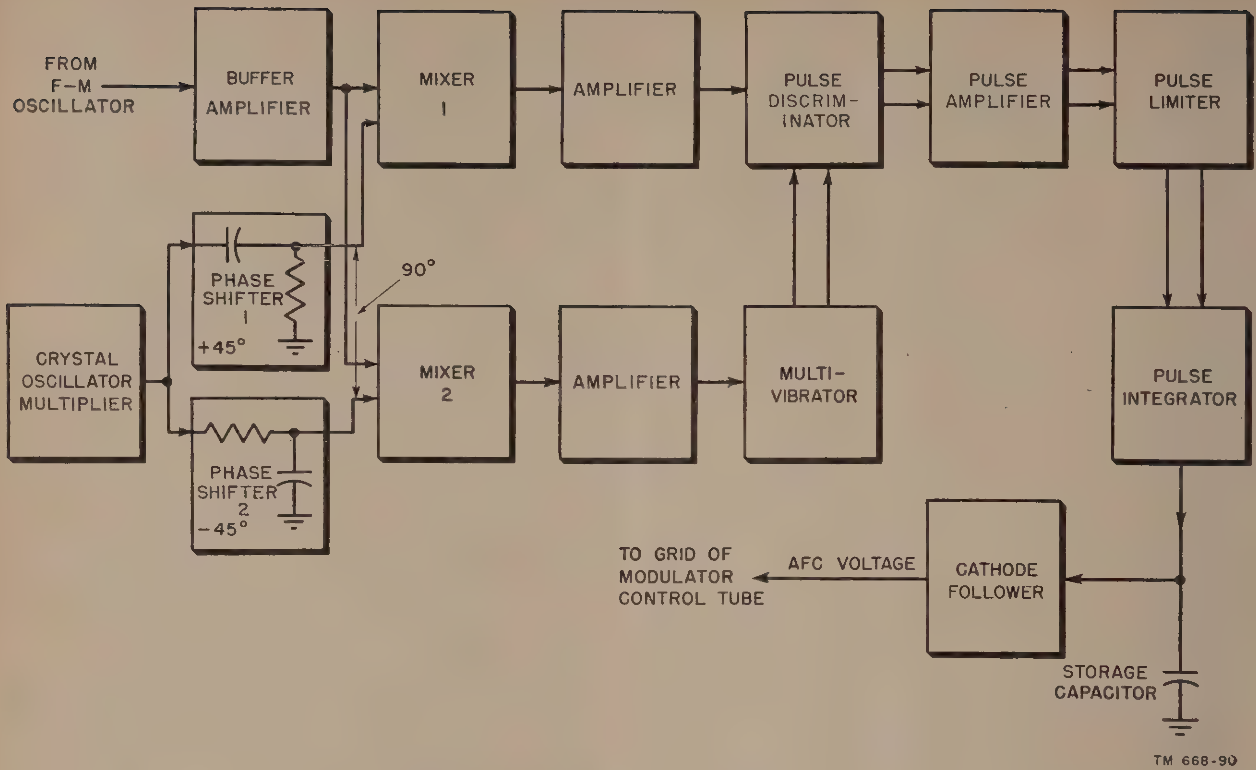


Figure 90. Afc system using pulse control.

that from the crystal oscillator are the same, and there is no output from the mixer because the difference frequency between the applied signals is zero and and the mixer is set to accept only the difference frequency. When the difference frequency is present, it is always very small compared to the frequency of either of the applied signals.

(5) When modulation is applied, the frequency of the master oscillator increases or decreases. Therefore, the input signal to the grids of the mixers will increase or decrease by a like amount. This signal mixes with that from the crystal oscillator and the difference frequency is equal to the instantaneous deviation of the mixer. When the modulated oscillator is on frequency, the upper and lower halves of each cycle of difference frequency contain the same average amount of area over a period of time. If the oscillator drifts upward, the output of the mixers is a wave in which average power in the upper half of the wave is larger than in the lower half. The control circuit is used to keep the average power in the upper and lower halves of the cycle equal by applying a suitable correction voltage to the master oscillator.

(6) Because of the phase-shifting networks, the output from the first mixer lags that of the second mixer by 90° when the frequency of the master oscillator is higher than that of the crystal. When the frequency of the oscillator is lower than that of the crystal, the output from the first mixer leads that of the second. At all times, the output of both mixers is 90° out of phase because of the constant difference in phase at the output of the crystal phase-shifting networks. During 1 cycle of modulation, the frequency of the oscillator first increases and then decreases. Therefore, the output of the mixers varies from minus 90° to plus 90°. This happens on each half-cycle.

That is, the phase lags on the high half-cycle of modulation and leads on the low half-cycle (fig. 91); f_o is the frequency of the oscillator and f_c is the center frequency.

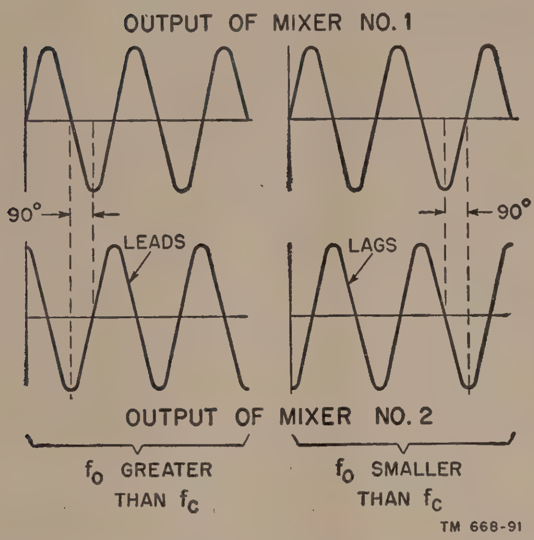


Figure 91. Phase relations in balanced modulators used in pulse control systems.

(7) The output from each mixer is passed through a separate amplifier. One amplifier is biased so that a few volts of negative signal drive it completely into cut-off. Therefore, the plate voltage increases to the value of the supply voltage and stays that way until the voltage on the grid allows the tube to conduct. When the grid swings positive, it draws current, and drives the tube to saturation. The plate current cannot increase and the plate voltage stays constant until the voltage on the grid goes negative. The result is the production of a square wave. The output from the other amplifier is used to trigger a multivibrator circuit that also develops a square wave output. This circuit acts as a switch and turns the d-c supply voltage off and on in response to the input signal. Each positive half-cycle develops a corresponding pulse in the output.

(8) The multivibrator circuit produces two square waves that are 180° out of phase with each other. The out-of-phase square waves are passed through R-C differentiating networks, as shown in figure 92. These differentiating networks are chosen so that the capacitance is small compared to the resistance. Therefore, the reactance of the capacitor is large at low frequencies and only high frequencies are passed. The square wave can be considered as composed of a number of sine waves of various frequencies and phase relationships. The highest of these occur at the leading and trailing edges of the wave. Therefore, the R-C networks produce pulses that are sharp and narrow.

(9) The pulses produced by the leading and trailing edges of the square waves are applied to the plates of a dual-diode pulse discriminator. The square waves from the first mixer and amplifier are applied to the junction of the two resistors. The cathode is biased negative to an amount equal to the peak value of the out-of-phase square waves. Therefore, neither the out-of-phase pulses nor the clipped signal from the first mixer can make the diodes draw current. However, when they are both present at once and in phase they are effectively in series at the diode plates.

(10) In figure 92, diode *D*1 is conducting because the voltages are adding in the wrong direction on *D*2. The output signal from *D*1 of the pulse discriminator is in the form of short positive pulses. Because of the bias in the output, only signal amplitudes larger than the bias can appear. Therefore, the actual output is not the square wave plus pulses, but the pulses alone. The discriminator produces positive pulses from *D*1 when the frequency of the master oscillator is above that of the crystal oscillator and negative pulses from *D*2 when it is below. The repetition rate of the pulses depends on the amount by which the two oscillators differ in frequency.

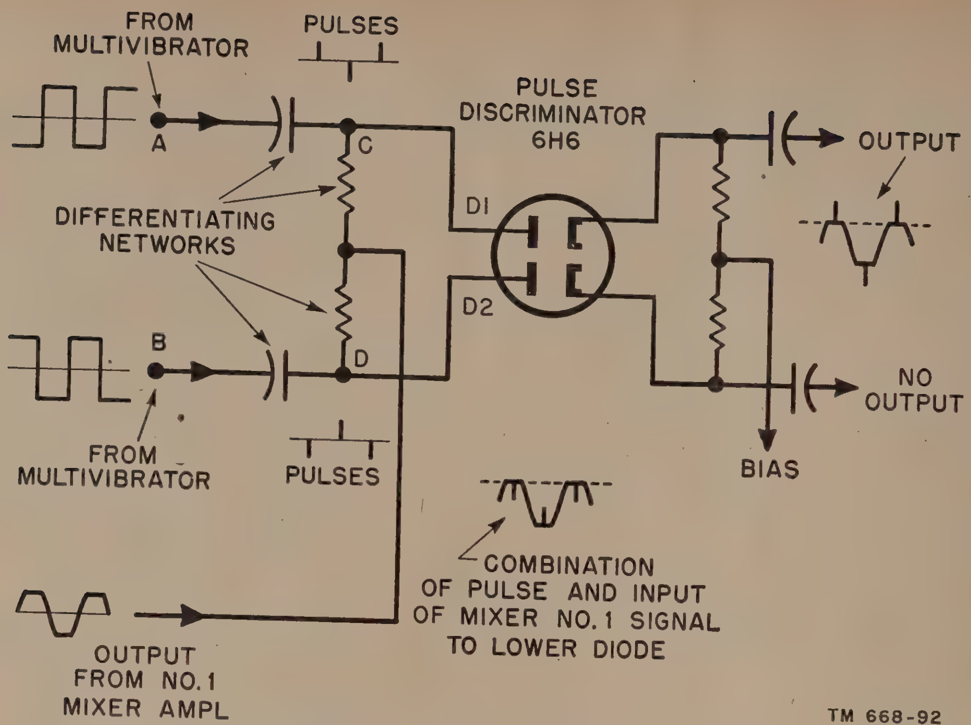


Figure 92. Pulse discriminator.

(11) The pulses are made uniform in amplitude by passing them through an amplifier and a pulse limiter, as shown in figure 90. They then are applied to a capacitor in the pulse integrator and each pulse charges the capacitor a small amount. When the frequency of the master oscillator swings high on modulation, more positive pulses are produced, and the capacitor charges to a positive voltage. On the downward swing of the carrier the reverse is true, and the capacitor charges negatively. If the upper and lower swings are equal about the constant value of the crystal oscillator, the net charge across the capacitor is zero. However, when the oscillator drifts higher or lower, there is an increase in the number of negative or positive pulses. This alters the net charge on the capacitor and is equivalent to a changing voltage across it. This changing voltage is applied to the grid of the modulator tube to correct any drift in the oscillator itself.

(12) This system needs no tuned circuits, and stabilizes on alternate halves of the modulation swing. The response time of the system to changes in center frequency can be as low as the period of 1 cycle at the lowest audio frequency. Long-time variations also are compensated for, since these depend only on the number of pulses and not on changes in supply voltage or tubes. The over-all circuit can be used only in large fixed transmitters where the number of component parts is not a problem.

44. Motor Control Systems

a. Description. The motor control system uses a two-phase motor attached to a variable capacitor and connected across the master oscillator tank. The frequency-control circuit derives two voltages 90° out of phase to turn the motor if the frequency of the master oscillator does not coincide with the frequency of the crystal oscillator.

b. Circuit. The frequency-control section of a motor-control system consists of three main parts—the crystal oscillator, the frequency di-

vider, and the motor-control section. Some of the signal from the master oscillator is tapped off and fed to a chain of frequency dividers as shown in the block diagram of figure 93. The low-frequency divided signal is fed to two balanced-modulator circuits. The output of the crystal oscillator section is divided, so that the frequency is the same as that of the divided signal from the master oscillator. It then is split up, shifted in phase by 45° in each half, and applied to the balanced modulators. The output of the balanced modulators is fed to a four-winding two-phase motor. When the phase in all windings is the same, there is no rotation of the motor. If the master oscillator drifts, the phase in two of the windings differs from that in the other pair, and the motor rotates to correct the unbalance.

c. Balanced Modulator Motor Control System. The output of the crystal-oscillator divider system is split up by a phase-shifting network and applied to the grids of a balanced modulator. The input from the master-oscillator divider system also is applied to the modulator grids. When the two frequencies are equal, no output appears in the plate circuits of the modulator, but any difference between the frequencies of the input signals will cause an output to appear. This output is an a-c voltage with a frequency equal to the difference in frequency between the two input signals. Each balanced modulator output is connected to two windings of a four-winding, two-phase motor. When the frequencies of the crystal oscillator and the master oscillator in the modulators coincide, the difference frequency in the output is zero, and no voltage is applied to the motor. When a difference frequency is present, the output of one balanced modulator is 90° out of phase with the other. This 90° out-of-phase pair of voltages applied to the motor windings causes a rotating magnetic field to be set up. Therefore, the armature of the motor turns with speed of rotation equal to the speed of rotation of the magnetic field. This, is turn, depends on the frequency of the difference of the signals applied to the modulators. The motor, in turning, moves the capacitor connected across the oscil-

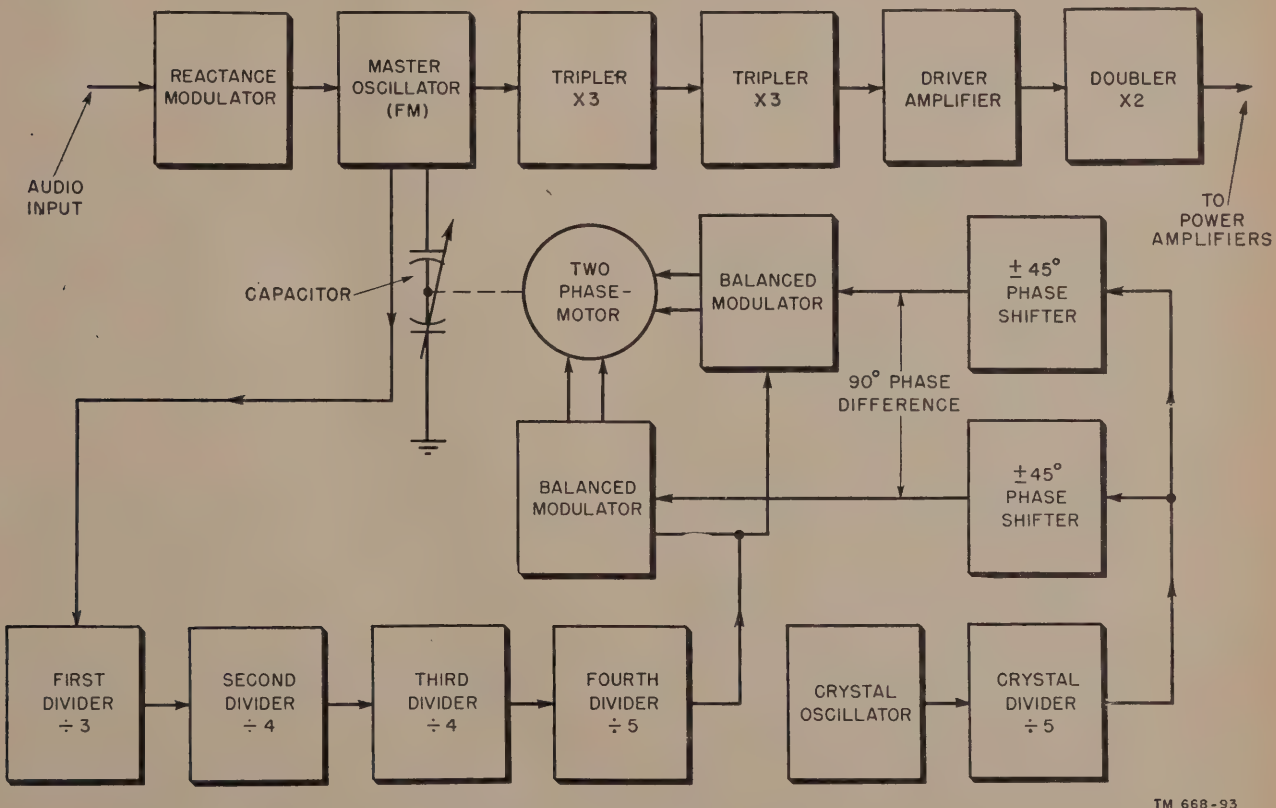


Figure 93. A motor-control afc system.

lator tank, so that the frequency of the master oscillator also changes. As the frequency changes, the difference between the master oscillator and the crystal frequency becomes less and less and the motor runs more slowly until it stops when the two frequencies coincide. Therefore, a small displacement of the oscillator frequency causes the motor to turn a very small amount and correct the drift.

45. Frequency-Divider Circuits

a. General. Frequency-divider circuits reduce the frequency of a signal by an integral multiple of the fundamental and also can divide and multiply by fractional quantities. They also reduce any deviation that is present on the signal because of modulation. There are two general types of frequency dividers, those that produce an output whether an input signal is present or not, and those which produce output only when the input signal is applied. The free-running dividers are all some variety of oscillator synchronized with a higher frequency. The other type of divider depends on the properties of special circuits and not only can produce division and multiplication, but also can divide and multiply by fractional quantities.

b. Multivibrator.

(1) One of the simplest oscillators that can be used as a frequency divider is the synchronized multivibrator. There are many varieties of multivibrator circuits, but essentially they are all modifications of a two-stage resistance-coupled amplifier circuit with the output fed back to the input circuit. When the grid voltage of a vacuum tube is made more positive the plate voltage decreases. This decrease in plate voltage is coupled into the grid of one tube, causing a decrease in grid voltage. This results in an increase in plate voltage, which is applied to the grid of a second tube, and the cycle reverses. The circuit is shown in figure 94. The variations possible consist in using direct coupling, cathode coupling, or mixed types of coupling between the two tubes.

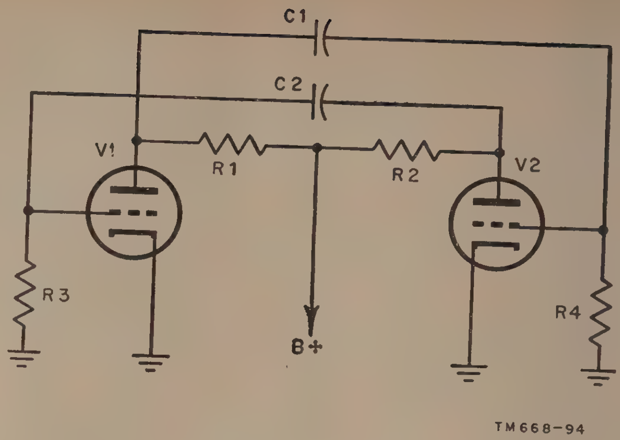


Figure 94. Simple multivibrator circuit.

(2) A small amount of voltage applied to the grid circuit can be used to trigger oscillation. Any voltage that is an integral multiple of the natural frequency of the oscillator provides this triggering action. The frequency can be much higher than the actual frequency of operation of the oscillator. The output from one multivibrator controlled in this manner can be ten times less in frequency than the controlling voltage. The output of this multivibrator can be connected to another multivibrator that also divides by a like amount, providing division by one hundred. In this way, the high frequency of the crystal oscillator and the master oscillator in an f-m system can be reduced to a frequency in the audio range. This can be applied to the phase or pulse discriminators for frequency control, as described previously.

(3) If the synchronizing voltage is not applied to the multivibrator, the oscillations do not stop, but run freely; hence, the name, *free running*. This is a distinct disadvantage in a divider circuit because if synchronism is lost, frequency control also is lost. For this reason a modification of the multivibrator known as the *one-shot* multivibrator or *trigger circuit* often is used (fig. 95). The circuit is essentially the same as the multivibrator, but

capacitors $C1$ and $C2$ of figure 94 are replaced by resistors $R5$ and $R6$. It sometimes is called a direct-coupled multivibrator. A small change in grid voltage of $V1$ increases the plate current. This increases the voltage drop across $R1$ and makes the grid of $V1_a$ more negative, decreasing the plate current through $R2$. As the grid of $V1$ becomes more positive, there is an abrupt increase of the plate current of the first tube and the plate current of $V1_a$ is cut off. Another pulse applied to the grid of tube 2 upsets this condition and causes another reversal, with maximum current in $V1_a$. Therefore, this circuit depends entirely on the input pulse, and does not operate (there is no output) when no pulse is present.

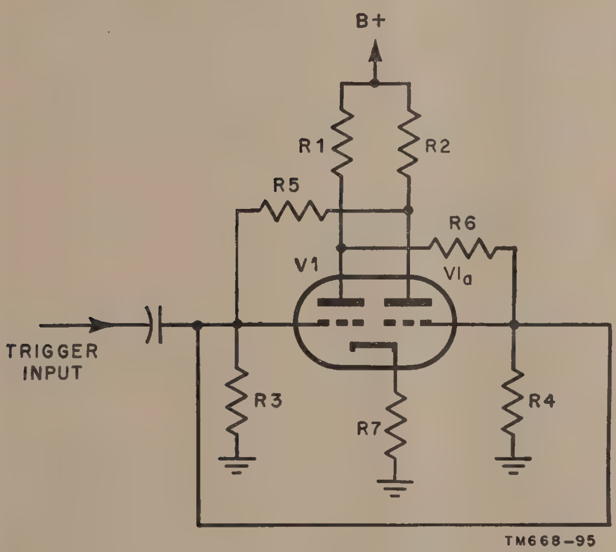


Figure 95. Trigger circuit.

(4) If several of the circuits described above are connected in sequence, the application of a pulse to the input of the first tube which was not previously conducting, conducts, and vice versa. The next stage is made to trigger only on the output of one conducting tube, and, since its plate voltage is low when conducting, the second stage does not trigger on the first pulse. A second pulse, applied to the first tube, triggers the second tube and causes it to conduct. A third pulse triggers the next tube in line and so on. If ten tubes are used, the tenth tube triggers on the tenth pulse. If this tube connects to a similar group of circuits, the last tube in the second line will trigger on every hundredth input pulse. This provides an accurate frequency divider that can be extended to very high orders of division. It has the advantage that there is no output unless an input signal pulse is present. On the other hand, a very large number of tubes becomes necessary if high division is needed.

c. Synchronized Oscillator. A vacuum-tube oscillator tends to synchronize with an injected voltage of about the same frequency. Also, if the frequency of the oscillator and the injected voltage are in approximate harmonic relationship, the oscillator synchronizes with the harmonic. For example, if a signal of 1 mc is injected into an oscillator operating at about 99 kc, the oscillator begins to oscillate at 100 kc, which gives a frequency division of ten. As the frequency stability of the oscillator is reduced, synchronization can be obtained over a wider and wider range. Synchronization can be improved if afc is applied to the oscillator. This is accomplished by applying the output of the oscillator and the voltage to be divided to a phase discriminator. The rectified output of the discriminator actuates a reactance modulator, which in turn changes the frequency of the oscillator. If a harmonic generator is inserted between the oscillator and the detector, the oscillator can be held in synchronism with its own harmonic, and therefore can act as an accurate divider. However, the circuit continues to function even with no synchronizing signal present. The oscillator therefore is free-running (fig. 96).

d. Regenerative Modulator. When two different frequencies are applied to a mixer circuit, the output contains all possible sum and difference frequencies and their harmonics. If one of the multiples of the difference is selected, amplified, and inserted back into the mixer circuit, it reinforces the output at that frequency. This is known as *regenerative modulation* (fig. 97). This device uses an ordinary

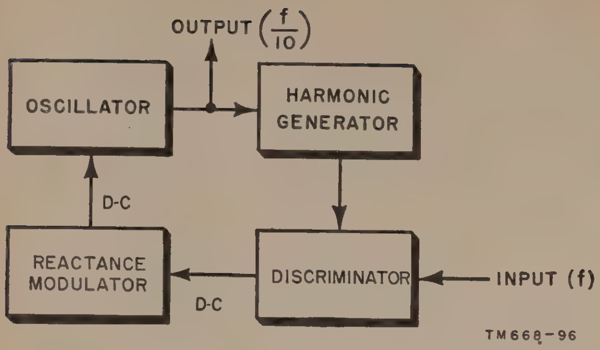


Figure 96. Synchronized-oscillator frequency divider.

mixer tube and a suitable selective amplifier. For example, the input to the mixer grid produces a distorted wave, which contains the tenth submultiple. The output is taken from the amplifier and fed back into the mixer, where it increases the amplitude of that submultiple and tends to suppress the others. The result is a divider that cannot operate unless an input signal is present. Fractional division can be obtained if the diagram of figure 98 is used. Here, the output of the amplifier is fed into an odd-harmonic generator, and an odd fractional value of the input signal is reinforced.

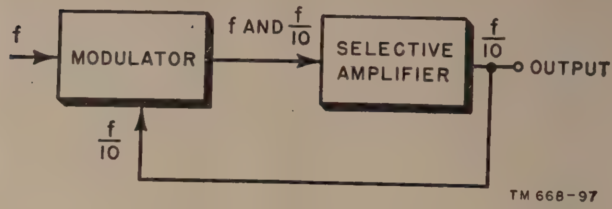


Figure 97. Regenerative-modulator frequency divider.

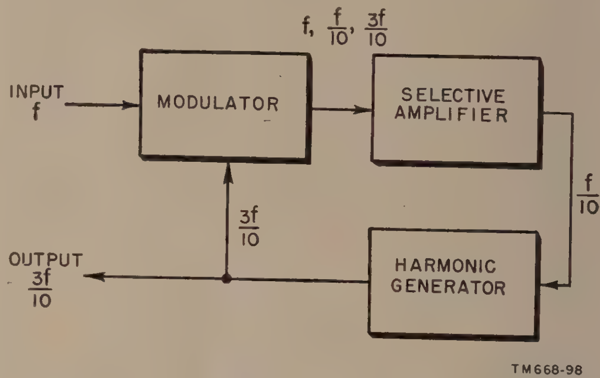


Figure 98. Regenerative-modulator fractional divider.

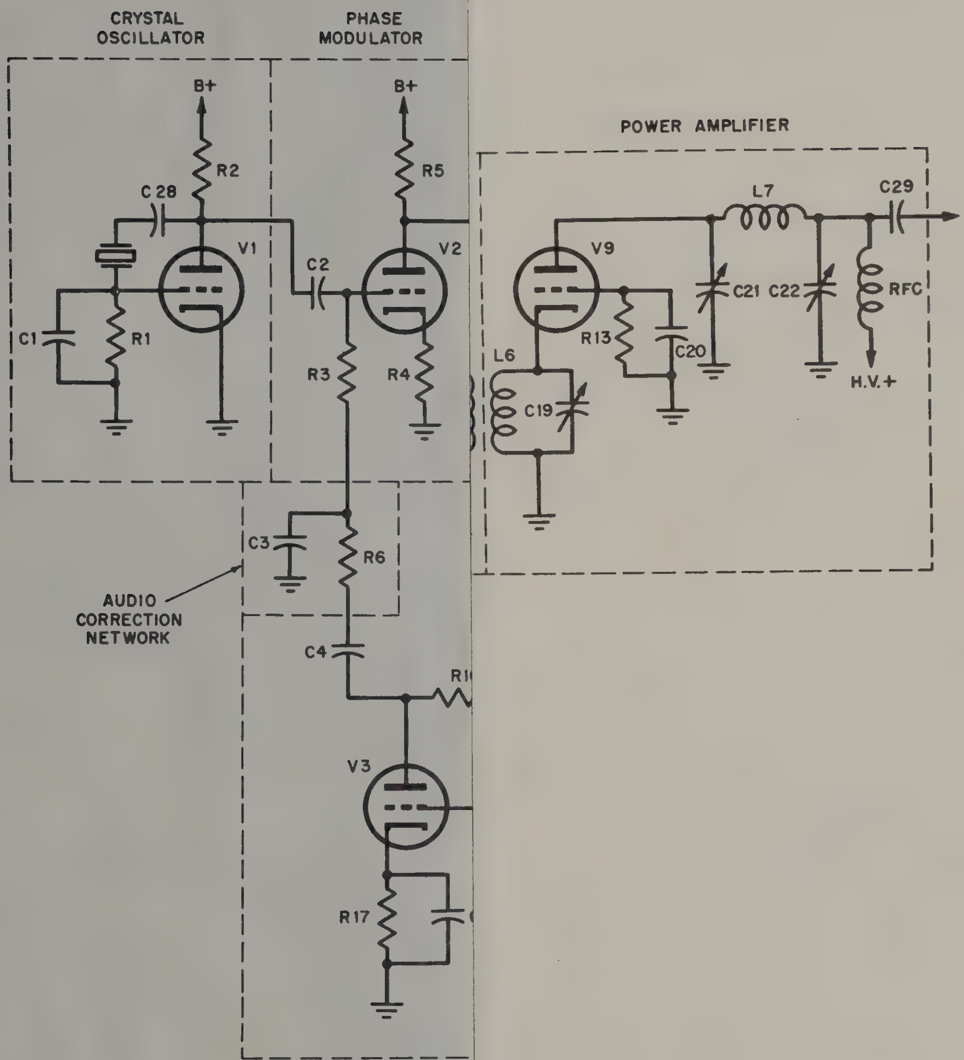
Section III. COMPLETE TRANSMITTERS

46. Indirect F-M Transmitter

a. The preceding sections of this chapter have discussed the operation of various circuits used in f-m transmitters. An over-all circuit diagram of these circuits combined in an actual piece of equipment would be large and complicated, especially if the transmitter has a large number of stages. An over-all schematic for a typical f-m transmitter, using indirect methods for producing the production modulation, is shown in figure 99. Although many things are shown at the same time, it is easy to follow the entire plan if the individual circuits that compose it are understood. For greater clarity, the power-supply wiring as well as the control circuits has been eliminated from this diagram. Normally, when printing the complete schematic, filament wiring, connections for plate voltage, and control circuits are all shown.

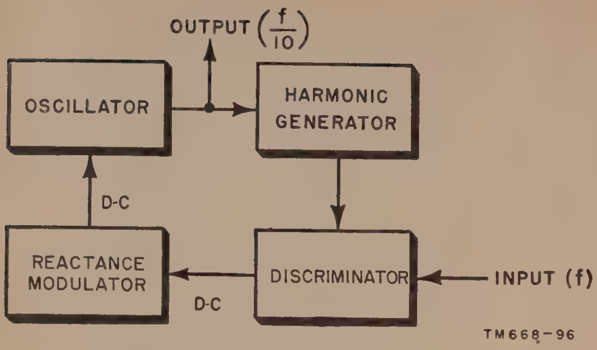
b. The indirect f-m transmitter has a Pierce crystal oscillator, V1, which is similar to the ultraudion oscillator. The crystal acts as a tuned parallel resonant circuit connected between the grid and the plate through a blocking capacitor, *C*28. Grid-leak bias is provided by the combination of resistor *R*1 and capacitor *C*1. Plate voltage is supplied through *R*2, and the output voltage of the oscillator is developed across it. The coupling capacitor, *C*2, couples the output voltage to the grid of the following stage.

c. The Link phase modulator is used in this stage, and the voltage drive for *V*2 is developed across grid resistor *R*3. Audio voltage is introduced through the correction network, *R*6 and *C*3, which produces the necessary frequency response required for the generation of true f-m. The high cathode bias that is used with this modulator is provided by *R*4, its value being great enough to operate the tube in a region of low transconductance. The frequency-modulated output is developed at the plate of *V*2.



CRYSTAL OSCILLATOR
PHASE MODULATOR
POWER AMPLIFIER
AUDIO CORRECTION NETWORK
B+
R2
C 28
V1
C1
R1
B+
R5
V2
C2
R3
R4
C3
R6
C4
V3
R17
V9
L7
C29
C21
C22
RFC
R13
C20
L6
C19
H.V.+

TM 668-99



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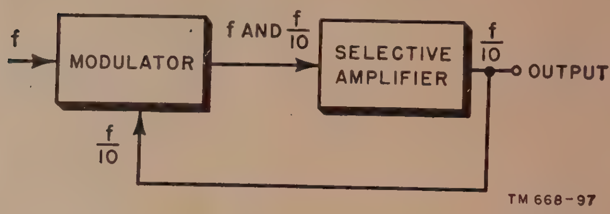
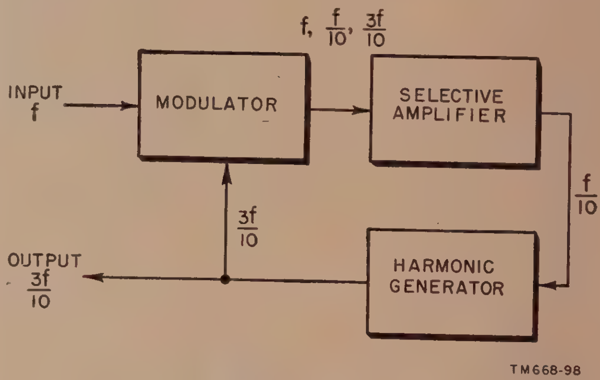


Figure 97. Regenerative-modulator frequency divider.



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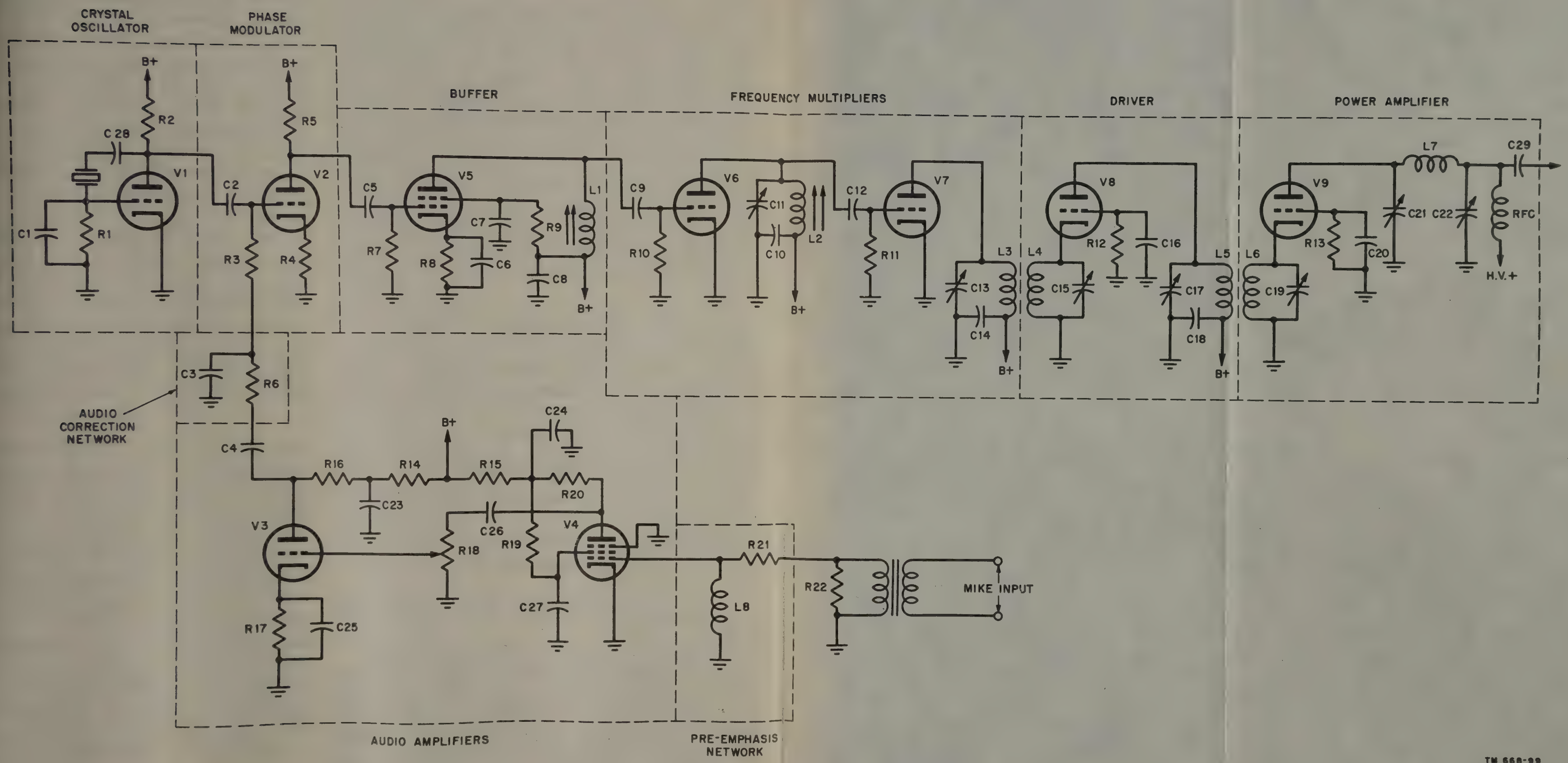


Figure 99. Indirect f-m transmitter.

d. The resistance-coupled a-f voltage amplifiers of the unit are formed by the circuits of *V*3 and *V*4. Included in the amplifier are a pre-emphasis circuit, a gain control, and decoupling circuits. The voltage developed by the microphone is isolated from the grid of the first audio tube, *V*4, by the microphone input transformer. At the same time this transformer provides a certain amount of voltage gain. Resistor *R*22 serves to stabilize the impedance of the secondary, so that the pre-emphasis network formed by *R*21 and *L*8 can function properly. Screen voltage and bypassing are provided by *R*19 and *C*27 respectively, and *R*20 is the conventional plate load resistor. Capacitor *C*24 in conjunction with resistor *R*15 acts as a decoupling network, to prevent feedback from the following stages from returning to the grid of *V*4 through the common power supply impedance. The same function is performed by *R*14 and *C*23 for the following stage. In this stage, *R*17 and *C*25 are the conventional cathode bypass and bias circuit, and *R*16 serves as the plate load resistance. Control of the voltage, fed from *V*3 through coupling capacitor *C*26 to the modulator, is achieved by variable resistor *R*18 in the grid circuit of *V*3. The output of the amplifier stages is supplied to the audio correction network, *R*6 and *C*3, through coupling capacitor *C*4.

e. From the modulator stage, *V*2, the frequency-modulated signal passes through capacitor *C*5 and builds up a voltage across *R*7 in the grid return of *V*5. This stage is a class A buffer amplifier that isolates the modulator and associated circuits from the frequency multiplier circuits which follow. It uses cathode bias provided by resistor *R*8 with bypass capacitor *C*6. The output voltage is developed across the tuned plate circuit formed by *L*1 and the distributed capacitance. Screen voltage is applied through *R*9, with *C*7 and *C*8 operating as conventional r-f bypass capacitors. The voltage developed across *L*1 is still at the crystal frequency, but has been increased considerably in amplitude.

f. The voltage across *L*1 is coupled to the grid of the first frequency multiplier, *V*6, through *C*9, and is rectified between the grid and cathode, since its positive swings are sufficient to draw grid current. This develops bias across *R*10, causing the stage to operate in class *C*. The desired harmonic is selected by the tuned plate circuit formed by *C*11 and *L*2, in which *C*10 serves as an r-f bypass capacitor. This higher frequency is coupled to another frequency multiplier, *V*7, identical in operation with *V*6. The components of the stage perform functions equivalent to those of the preceding stage. The tuned plate output circuit of the second multiplier, *V*7, is tuned to the operating frequency desired.

g. Drive from tuned circuit *L*3 and *C*13 is transformer-coupled to the cathode circuit of *V*8, *L*4, and *C*15. This stage supplies excitation for the power amplifier. Both the driver and the power amplifier are grounded-grid stages, as necessitated by the high operating frequency. Bias for the grounded grid is achieved through combination *R*12 and *C*16 in *V*8, and *R*13 and *C*20 in *V*9. Since sufficient voltage is applied to the cathode to make it negative in respect to the grid, current flows in the grid circuit and builds up bias across the resistor. The capacitor acts as a bypass, and effectively grounds the grid for r-f.

h. The plate output circuit of the driver stage formed by *L*5 and *C*17 is coupled to the power-amplifier cathode circuit, *L*6 and *C*19. Output from the entire transmitter is applied to a transmission line by the pi-network impedance-matching circuit formed by *C*21, *L*7, and *C*22. This circuit permits matching a wide range of transmission lines by adjustment of the variable capacitors. Plate voltage is fed on the low-impedance side of the circuit through an r-f choke, RFC, and is prevented from reaching the antenna by blocking capacitor *C*29.

47. Direct F-M Transmitter Circuit

a. The circuit for a direct f-m transmitter generally will be more complicated than most of the indirect types when it incorporates automatic frequency control. If no afc is used, considerable simplification in the number of stages can be obtained at the expense of reduced frequency stability. Because most direct f-m systems are capable of a higher deviation of the fundamental oscillator frequency than are indirect units, less frequency multiplication is needed to attain the desired operating frequency and deviation. A representative direct f-m transmitter with automatic frequency con-

trol is shown in the complete schematic of figure 100. Although the diagram looks extremely complex at first glance, it can be understood easily if it is analyzed stage by stage. The power supply and control circuits have been eliminated for clarity.

b. The circuit of tube *V*1 is a basic Colpitts oscillator. The frequency control network comprising *V*3 and *V*4, along with reactance modulator *V*2, serves to keep it in synchronism with crystal oscillator *V*5. Audio voltage is amplified by *V*13 and applied to the reactance modulator to produce the necessary deviation. The remaining stages in the transmitter are buffer amplifiers, frequency multipliers, and the final output stage.

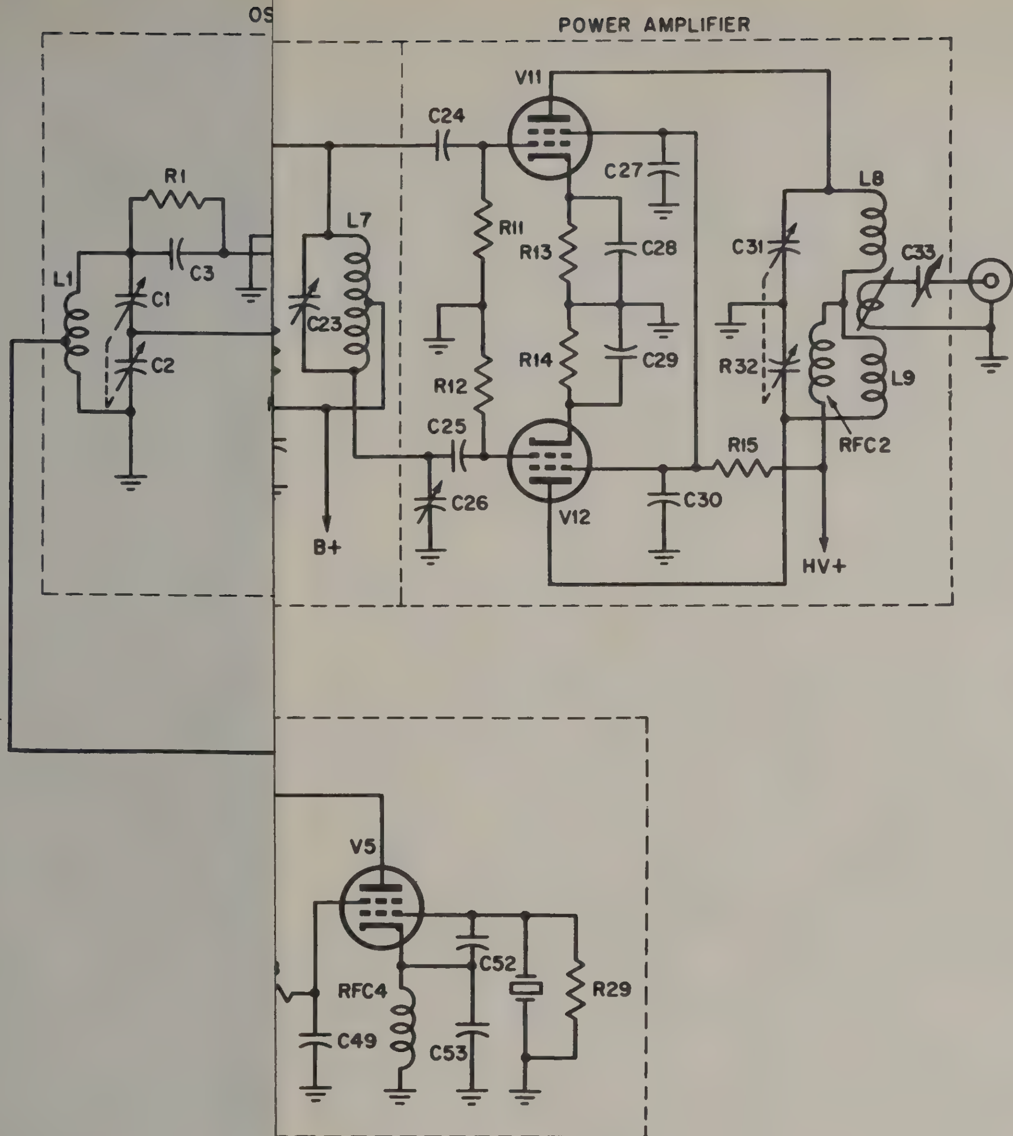
c. The tank circuit of oscillator *V*1 is formed by *L*1 and the split-feedback capacitors, *C*1 and *C*2. *R*1 and *C*3 are a conventional grid-leak bias combination. An r-f choke is used in the cathode so that d-c plate current can flow from plate to cathode, and r-f currents are diverted to the feedback circuit. A conventional plate tank circuit, *C*6 and *L*2, along with bypass *C*5, is used. This tank circuit is tuned to a harmonic of the oscillator frequency to obtain increased isolation of the frequency-determining components from the output of the oscillator. In addition, a stage of frequency multiplication is saved. Screen voltage is fed to the oscillator tube through the voltage divider formed by *R*2 and *R*37. When the screen voltage is adjusted properly, the oscillator frequency is practically independent of changes in plate voltage.

d. Output from the oscillator is coupled through *C*7 to the grid of buffer amplifier *V*6, which operates class A as determined by the bias built up across *R*4 in the cathode circuit. Capacitor *C*8 acts as a bypass for r-f and *R*3 serves as a d-c grid return. It does not affect the bias because the output of the oscillator is adjusted so that negligible grid current is drawn. The screen voltage is supplied from the main plate supply through *R*5. The screen itself is grounded for r-f by *C*9, and the bottom of the plate tank circuit, *L*3 and *C*58, is bypassed by *C*10. The output of this stage is fed to *V*7 of the first frequency multiplier through *C*11. Bias is developed by grid-cathode rectification of the amplified signal across *R*6. The output of this multiplier appears across the tuned circuit formed by *C*13 and *L*4, which is returned to ground for r-f by bypass *C*12. This output is fed in a similar manner to *V*8 of the second multiplier which in turn feeds *V*9 of the third multiplier. The output of the third multiplier appears across the tuned circuit formed by *C*18 and *L*6, which is bypassed to ground for r-f by *C*19.

e. Examination of the frequency modulator discloses that *V*2 is the conventional injected-reactance type, with *C*35 and resistor *R*18 forming the phase-splitting network. Resistor *R*17 is inserted in series with *C*35 to prevent the development of very-high-frequency oscillation, which can be caused by *C*35 in association with stray wiring inductance. This can act as a tuned circuit and permit the development of ultraudion oscillation at frequencies where the screen loses its effectiveness in reducing capacitance between grid and plate. *C*34 is a blocking capacitor that prevents d-c from appearing in the grid circuit. Isolation is increased by bypass capacitor *C*36. The screen voltage and bypassing are supplied by *R*20 and *C*37. Operating bias for the modulator is set by *R*19, and the cathode is grounded for r-f and audio by *C*38.

f. Audio voltage is fed to the modulator grid through isolating resistor *R*16 and coupling capacitor *C*54 from audio amplifier *V*13. The audio-amplifier stage has a pre-emphasis circuit in the grid formed by *C*57, *R*34, and *R*33. *C*57 is selected so that, in combination with *R*33, a voltage divider is formed which presents an increasing voltage with frequency caused by the decreasing reactance of *C*57 as the frequency is raised. To limit the attenuation of low frequencies *R*34 is included. Since it is in parallel with *C*57, the maximum impedance that can be developed is the resistance of *R*34 alone, when the frequency is so low that *C*57 has an extremely high impedance. *R*35 acts as a gain control and as a constant source impedance for the pre-emphasis network.

g. The frequency-control circuit includes *V*3, *V*4, and *V*5. It compares the output voltage of


OS
POWER AMPLIFIER
V11
C24
R1
L7
C27
L8
R11
C28
C31
C33
R13
C3
L1
C1
C23
R14
C29
R32
C2
R12
L9
C25
R15
RFC2
C26
C30
V12
B+
HV+
V5
C52
RFC4
R29
C49
C53

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f. Audio voltage is fed to the modulator grid through isolating resistor *R*16 and coupling capacitor *C*54 from audio amplifier *V*13. The audio-amplifier stage has a pre-emphasis circuit in the grid formed by *C*57, *R*34, and *R*33. *C*57 is selected so that, in combination with *R*33, a voltage divider is formed which presents an increasing voltage with frequency caused by the decreasing reactance of *C*57 as the frequency is raised. To limit the attenuation of low frequencies *R*34 is included. Since it is in parallel with *C*57, the maximum impedance that can be developed is the resistance of *R*34 alone, when the frequency is so low that *C*57 has an extremely high impedance. *R*35 acts as a gain control and as a constant source impedance for the pre-emphasis network.

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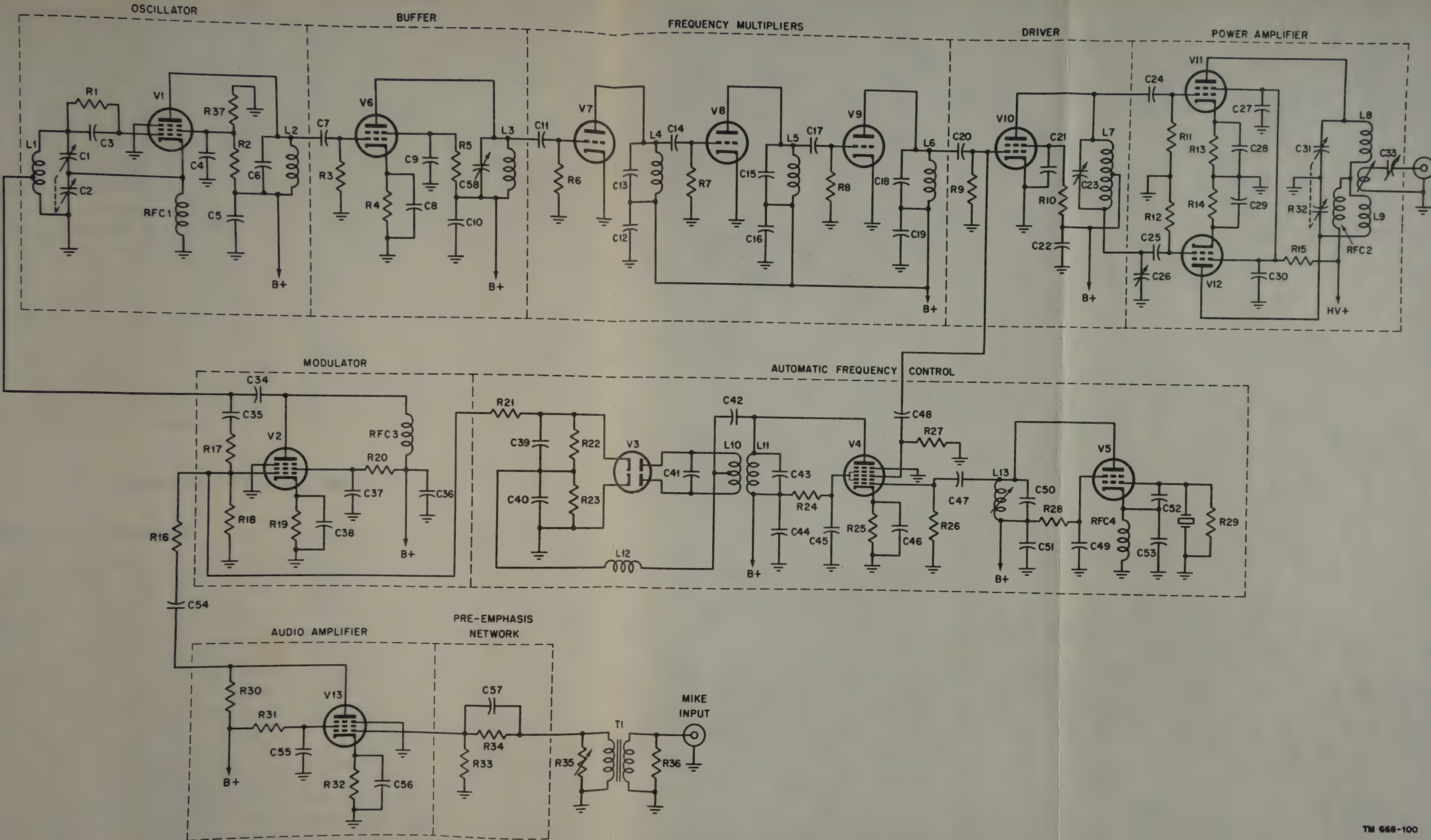


Figure 100. Direct f-m transmitter.

one of the multiplier stages with a harmonic of the crystal oscillator in a mixer, $V4$. If any difference exists between the two, a voltage is developed in the phase discriminator, $V3$, which is applied to the grid of the frequency modulator through isolation resistor, $R21$. This compensates for the frequency difference. The mixer stage uses a pentagrid tube which permits application to separate grids of the two signals to be compared. The difference frequency is selected by the plate circuit. This usually is tuned so that the entire discriminator operates at a low frequency. The crystal oscillator, $V5$, is a special type which permits the selection of a harmonic of the actual crystal frequency by means of the tuned plate circuit formed by $L13$ and $C50$. The harmonic of the crystal frequency is selected to give a difference between it and the harmonic of the master oscillator appearing at the grid of the driver tube, $V10$. $R29$ acts as the grid-bias resistor for the oscillator, with the crystal itself acting as the grid-leak capacitor. Output is coupled to the mixer tube through $C47$ and $R26$, and the output from the frequency multiplier $V9$ is coupled through $C48$ and $R27$.

h. The output of the frequency multiplier stages feeds a tetrode driver stage whose plate circuit is tuned to the operating frequency. This tuned circuit, $C23$ and $L7$, is grounded in the center for r-f by capacitor $C22$ so that a 180° out-of-phase voltage is available for the grids of the final push-pull amplifier, $V11$ and $V12$. A small variable capacitor, $C26$, is included on the side of the tank opposite the plate of the driver. It compensates for the capacitance from plate to cathode that would tend to unbalance the driving voltages to the grids of the power amplifier. Cathode bias is provided in the amplifier by $R13$ and $R14$ as a safety measure in case excitation fails, but the main operating bias comes from grid rectification and resultant production of voltage across $R11$ and $R12$. The final tank circuit, formed by $C31$, $C32$, $L8$, and $L9$, is coupled to a transmission line by the tuned link circuit. The screen-voltage supply for the final amplifier is conventional and is similar to that used for the driver.

Section IV. TYPICAL TRANSMITTERS

48. Indirect F-M Transmitter

A general description of a transmitter must include characteristics other than circuitry. These include the power input and output, the frequency and type of operation, the purpose of the equipment, and the distance over which the signal from the transmitter can be received. A description of these characteristics is given for the indirect f-m transmitter shown in figure 101. This equipment is designed for fixed-station service over a range of approximately 20 miles, with an output of 50 watts. A 110-volt, 60-cycle power supply is required for the power input, and the equipment draws 325 watts from the line when transmitting. The complete unit contains 28 tubes including rectifiers, and has an audio response from 300 to 4,000 cycles, with a frequency range from 30 to 40 mc. The transmitter unit has been removed from the cabinet and is shown in figure 102. The crystal oscillator is phase-modulated through an audio correction network. The resultant f-m signal is multiplied 32 times and produces a frequency deviation of ±15 kc. This is accomplished by 2 quadrupler stages and a doubler stage. The output of the doubler then is fed to push-pull power amplifiers, and approximately 50 watts of power is delivered to the antenna. The receiver (fig. 101) shown in the rack below the transmitter is a double conversion superheterodyne.

49. Direct F-M Transmitters

a. A combination double-conversion superheterodyne receiver and direct f-m transmitter is shown in figure 103. This set provides two-way phone communication with a similar portable, ground, or mobile equipments over a range up to a mile. Power is supplied by either dry cells, or a vibrator power supply and vehicular

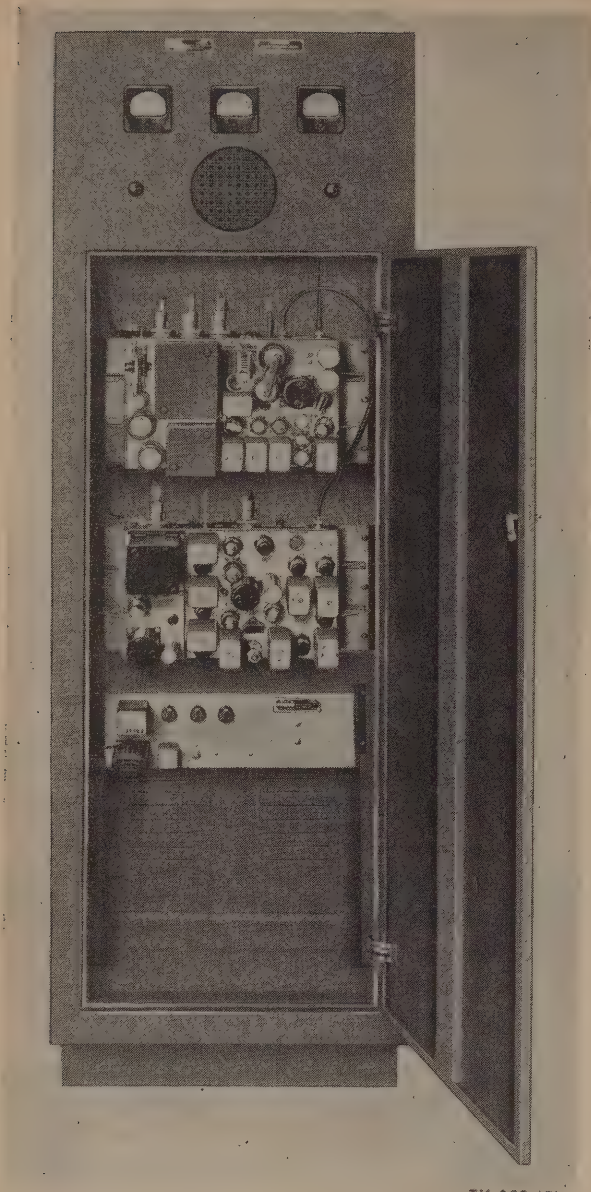


Figure 101. Indirect f-m transmitter used for fixed-station service.

battery. The power supply must provide 90 volts at 80 ma for the plates and screens of the tubes, and 6.3 volts at 520 ma when transmitting. The power output of the transmitter is 500 mw (milliwatts). The frequency deviation is ±20 kc when a 1,000-cycle signal input of .25 volt is applied. The tuning of the transmitter is variable continuously from 47 to 58.4 mc. This band of frequencies provides 115 channels with each channel having a bandwidth of 100 kc. Two preset detented channels are available. The receiver i-f components and the audio components of both the transmitter and receiver are mounted on the i-f chassis, as shown on the right side of figure 104. The view on the left of this figure shows both r-f and i-f components.

b. The transmitting and receiving circuits are associated with each other through a common antenna circuit, a common 32- to 42.3-mc Colpitts oscillator circuit, and a common tuning control. The transmitter converts speech signals from an external microphone, amplifier, telephone line, or other a-f source into f-m signals. The microphone voltage is amplified by a microphone amplifier to the proper value for modulation. It then is applied to a reactance modulator which varies the frequency of the Colpitts oscillator in accordance with the audio signal. The frequency-modulated output of the Colpitts oscillator and a 15-mc signal produced by doubling the low-frequency output of a 7.5-kc crystal oscillator are combined in a mixer stage and the sum frequency of the transmitter mixer is selected by a tuned circuit. The signal then is fed through the driver and power amplifier stages to the antenna. No antenna switching is provided since the receiver is inoperative when the transmitter is energized and the transmitter is inoperative when the receiver is on.

c. Another direct f-m transmitter receiver for portable, ground, or vehicular installation is shown in figure 105. The frequency range of this transmitter receiver is from 20 to 27.9 mc. The communication range is approximately 10 miles for vehicles in motion and 15 miles for stationary vehicles. Power can be supplied by dry cells, vibrator power supplies, or a hand generator. The set contains two subchassis; the i-f chassis mounts the receiver components and the audio components for both transmitter and receiver; the r-f chassis contains the high-frequency parts of both transmitter and receiver. A low-frequency, self-excited, modified Hartley oscillator is modulated by a reactance tube to produce direct f-m. A crystal oscillator controls the frequency of both the transmitter and the

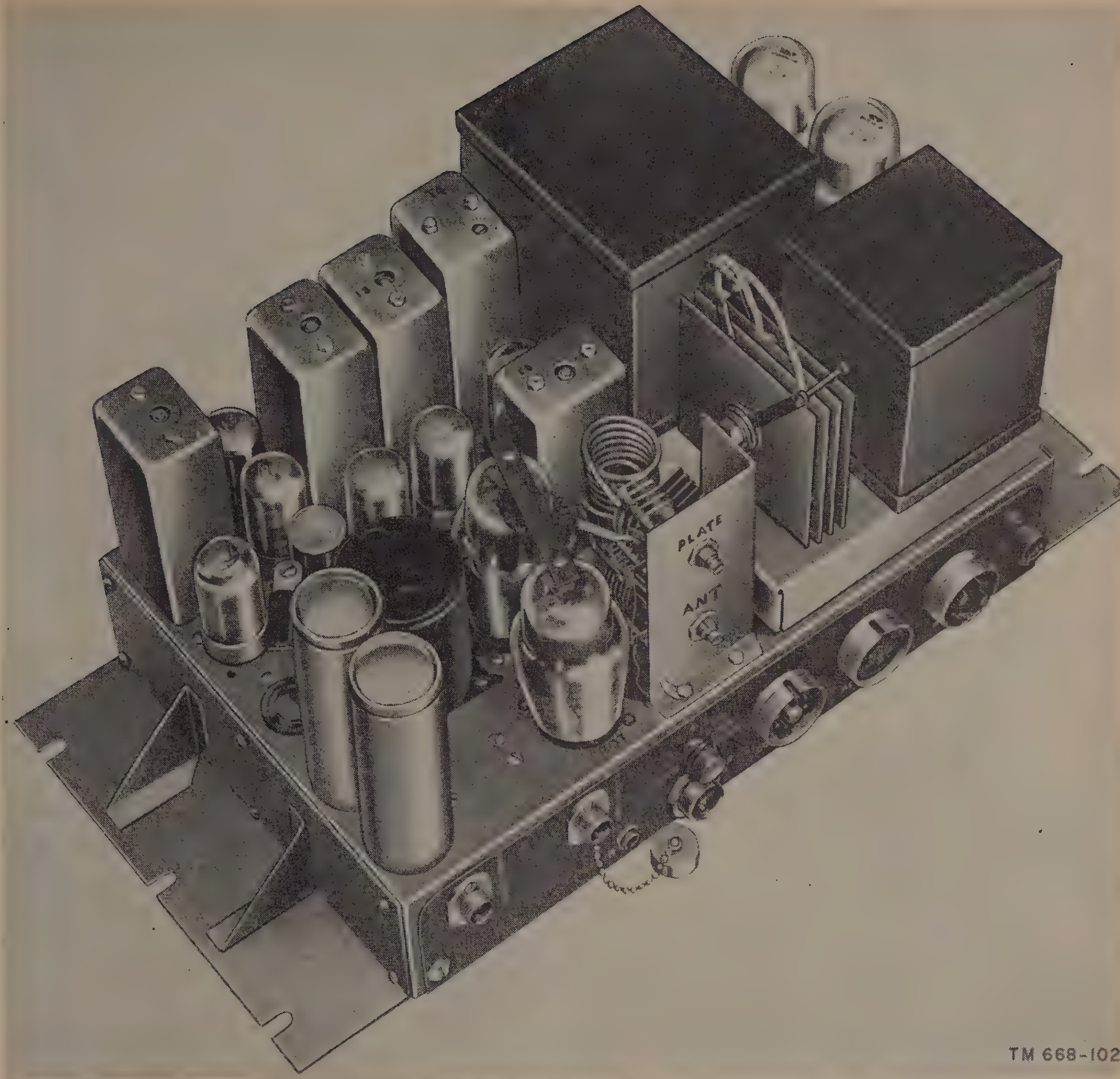


Figure 102. Indirect f-m transmitter removed from rack.

receiver. The output of this oscillator is mixed with the f-m signal to produce the desired deviation of ± 20 kc. The output of the mixer then is fed through the driver and power amplifier stages to the antenna.

d. A direct f-m receiver transmitter unit designed for pack operation is shown in figure 106. This set has an average distance range of 5 miles. The frequency range is from 27 to 38.9 mc. This band of frequencies provides 120 chan-

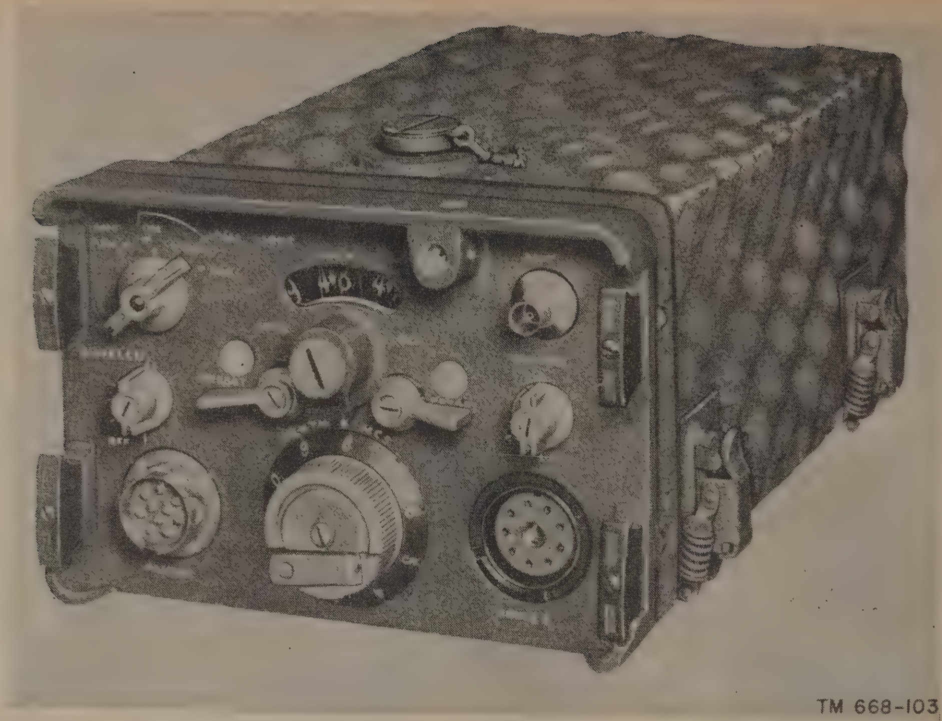


Figure 103. Direct f-m receiver-transmitter.

nels 100 kc wide and the transmitter is preset to any two of these channels. Power can be supplied by dry cells or storage batteries and a dynamotor. The power output of the transmitter is 1½ watts. The entire unit contains 19 tubes, all of which are used when transmitting. The transmitter is a MOPA (master oscillator power-amplifier) with the receiver crystal controlling the frequency of transmission through a reactance tube.

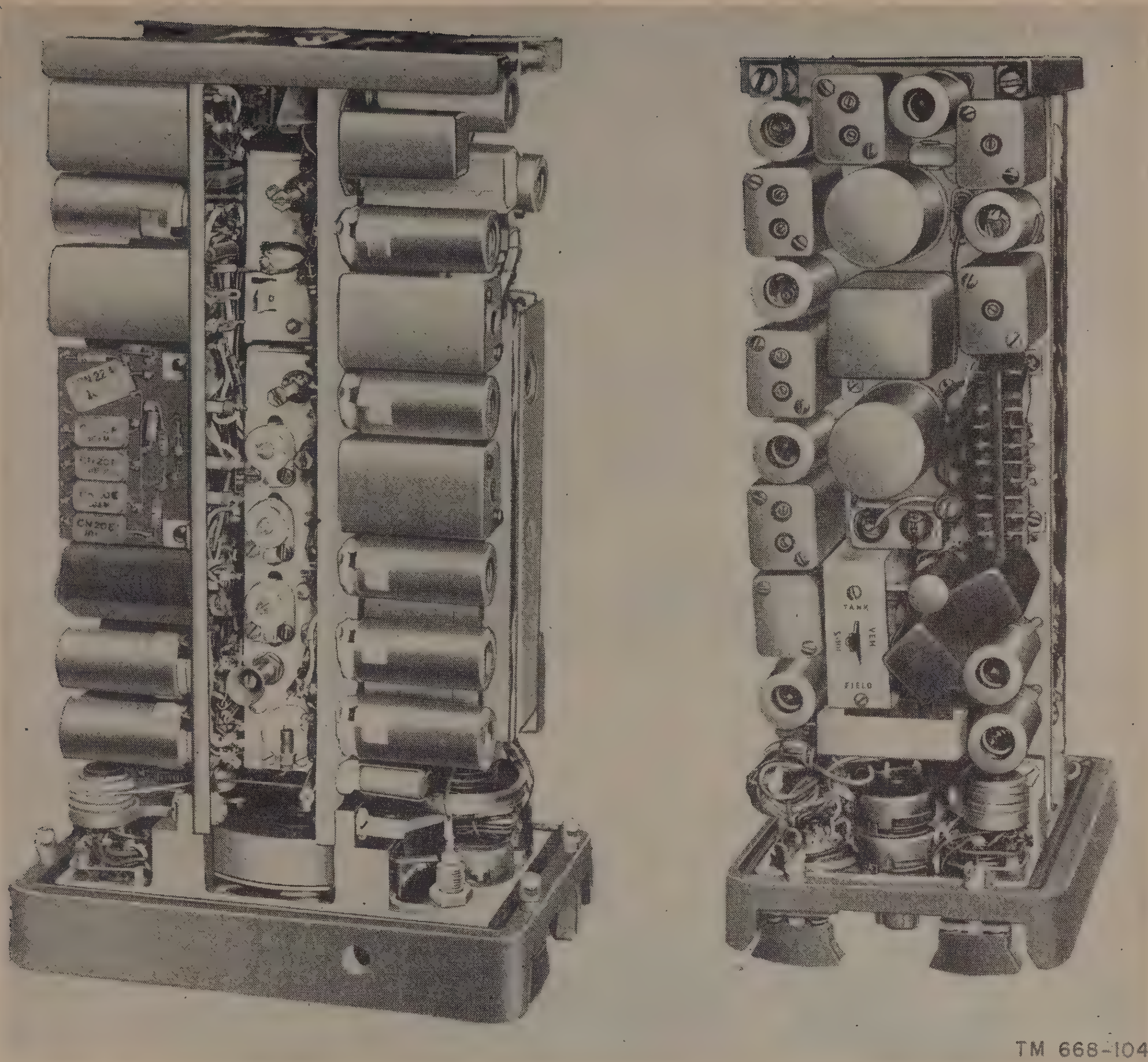


Figure 104. Direct f-m transmitter-receiver subchassis.

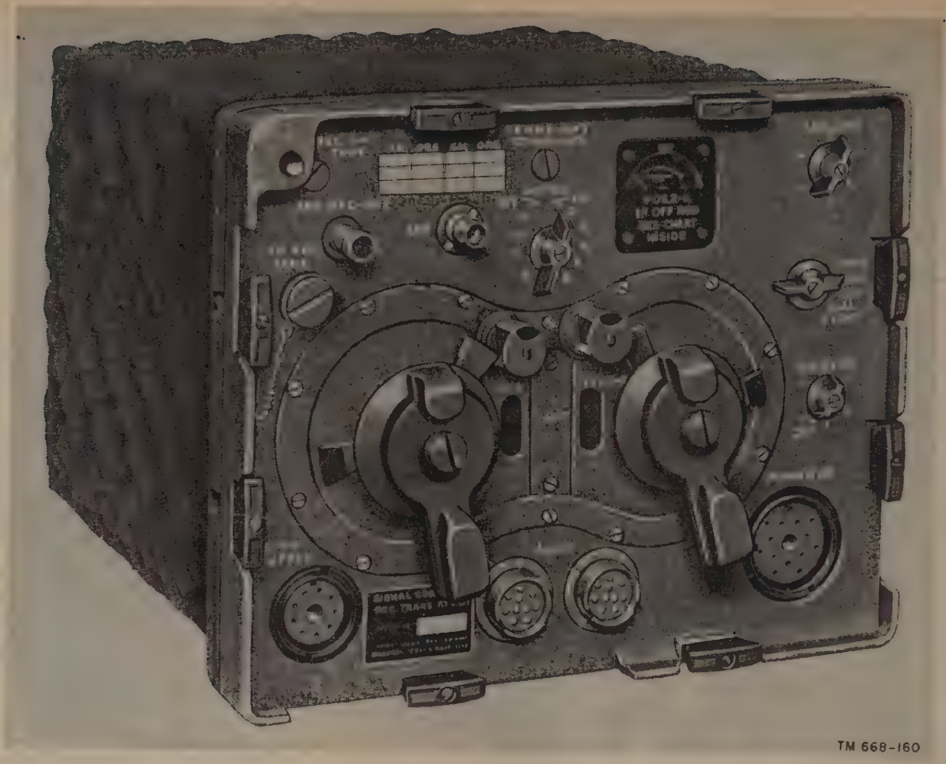


Figure 105. Direct f-m transmitter for vehicular use.

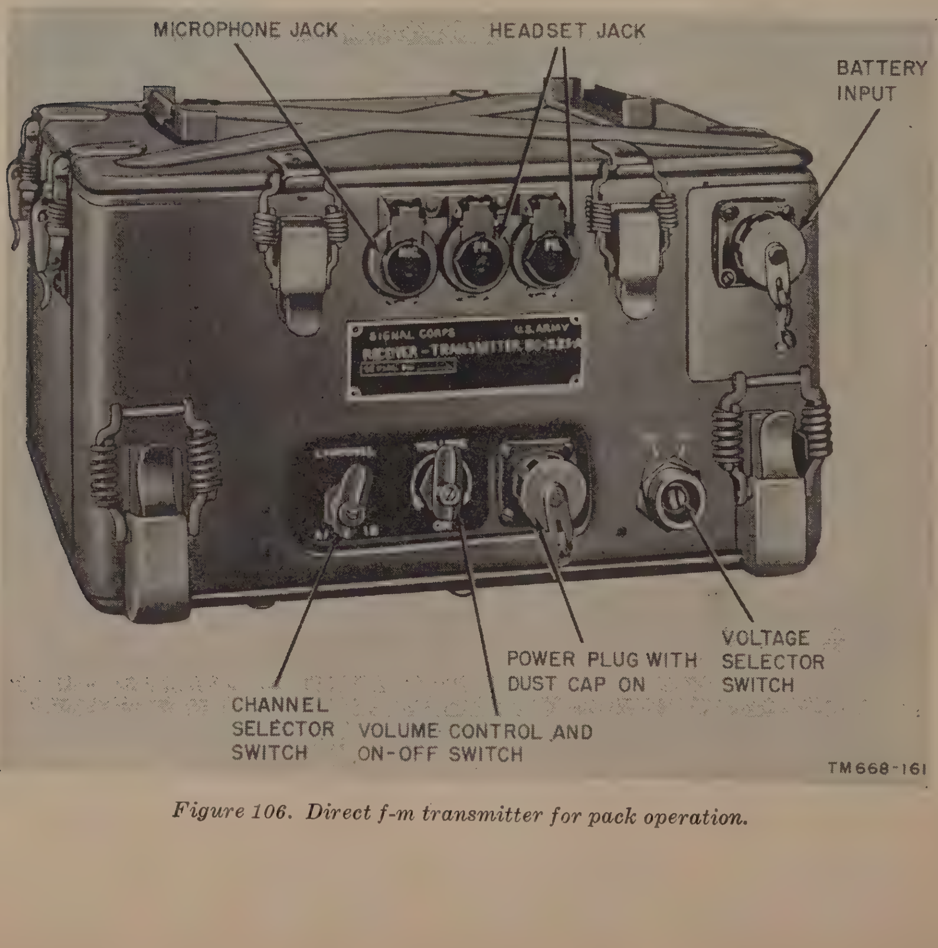


Figure 106. Direct f-m transmitter for pack operation.

50. Summary

a. The f-m signal taken from the modulator-oscillator stage must be increased in deviation and frequency by frequency-multiplier stages.

b. Frequency multipliers operate through harmonic generation. Multiplications of two, three, four, and five times are obtainable with reasonable efficiency.

c. Frequency multipliers are usually r-f amplifiers with input and output circuits tuned to different frequencies harmonically related to each other.

d. Push-push doublers are very efficient multipliers which use two tubes with grids connected in push-pull and plates in parallel. Operation is similar to a full-wave rectifier producing a ripple voltage of twice the line frequency.

e. At high frequencies, the operation of frequency multipliers is impaired because of degenerative effects which tend to reduce their power output. These degenerative effects can be neutralized with appropriate circuits.

f. Other methods of frequency multiplication are available which do not depend directly on harmonic generation, but these seldom are used.

g. Frequency multipliers can be combined with the master oscillator circuit for compactness and reduction of power consumption.

h. Power amplifiers for f-m have no connection with the modulation process. They are used only to increase the power in the modulated wave. All f-m power amplifiers operate class C, which permits high efficiency and high power output.

i. The operation of a class C amplifier depends on the angle of plate current flow, which in turn depends on the grid bias and the amplitude of the grid-driving voltage.

j. Various class C amplifier circuits have been devised with different means of introducing the grid-driving voltage and coupling the output power to the load.

k. The necessity for neutralization is overcome at high frequencies through the use of tetrode amplifiers.

l. Grounded-grid triode amplifiers overcome some of the disadvantages of tetrodes and conventional triodes at very-high frequencies.

m. The input circuit for a power amplifier must provide adequate regulation of the driving voltage without causing excessive losses. Various input circuits have been devised by use of capacitive coupling or inductive coupling.

n. Power-amplifier output-coupling circuits are of two forms, series-fed and shunt-fed. The output load circuit matches the final amplifier to the transmission line or the antenna. In addition, it suppresses spurious frequencies. Various types of output circuits provide different degrees of suppression and ability to match loads of different impedance.

o. Not all of the various input and output circuits available for class C power amplifiers can be used with each other because of the possibility of spurious oscillations being generated in the circuits.

p. Spurious oscillation generated in a power amplifier at a frequency far from the actual tuning of the tank circuits is called parasitic oscillation. Parasitic oscillations are eliminated by the proper choice of circuit in both grid and plate, as well as by suppressor chokes, resistors, and capacitors.

q. In direct f-m transmitters, the center frequency must be stabilized by some form of automatic frequency control.

r. There are two major types of frequency-control arrangements. One uses a d-c correction voltage for the modulator circuit, the other a mechanical arrangement.

s. Afc systems include frequency changers, comparators, and corrector networks.

t. The comparator is usually a discriminator circuit. The discriminator produces a d-c voltage proportional to the difference between the frequency of the oscillator and some standard center frequency.

u. The double-tuned discriminator uses two resonant circuits separated slightly in frequency and coupled to the same source. The output is applied to rectifiers and the amplitude of the rectified voltage is proportionate to the change in frequency.

v. The phase discriminator compares the phase of signals from the master oscillator and a crystal standard, producing a d-c output proportional to the difference.

w. The pulse discriminator produces pulses whose polarity and number per second reflect the departure of the oscillator from the center frequency. These pulses are stored, and any change in stored charge is applied to the modulator tube as a correction voltage.

x. In a motor control system, two balanced modulators, receiving their signals from the crystal oscillator and master oscillator, produce two outputs 90° out of phase if the signal frequencies do not coincide. The 90° out-of-phase voltages are applied to the windings of a two-phase motor. This motor turns a variable capacitor in the oscillator circuit, which serves to correct the frequency difference.

y. Frequency dividers are of two types—free-running dividers which produce output whether input is present or not, and those which produce an output only with an input signal.

z. The multivibrator is a common free-running divider. It is essentially a two-stage resistance-coupled amplifier with some of the output returned to the input. The essentially square waveform it produces can be synchronized with higher harmonics.

aa. Trigger circuits are modifications of multivibrator circuits which operate only in the presence of input pulses. Used in sequence, they operate as very accurate dividers.

ab. The regenerative modulator consists of a nonlinear device and a frequency selective amplifier. The amplifier is tuned to a subharmonic of the input frequency of the nonlinear mixer. The output of the amplifier is returned to the mixer, reinforcing the subharmonic.

51. Questions

a. What increases the deviation and carrier frequency in an f-m transmitter?

b. What feature of the operation of a class C amplifier permits the successful generation of output power at a harmonic frequency of the input?

c. Cite two ways of obtaining a frequency multiplication of 32.

d. How many times is the deviation of a basic f-m signal increased after passing through four doublers?

e. If the input circuit of a quintupler is tuned to 10 mc, to what frequency is the output circuit tuned?

f. Which type of multiplier chain produces greater output, one containing a doubler and a quadrupler or one containing three doublers?

g. How are the grids and plates of a push-push doubler connected?

h. What type of loading is produced by the capacitance between the grid and the plate circuit of a high-frequency multiplier?

i. What is the effect of a high-cathode inductance on the output capacitance of a high-frequency quadrupler?

j. What is the advantage of using push-pull multipliers?

k. What coupling exists between the input and the output circuit of a frequency multiplier which is combined with an oscillator?

l. Why is it desirable to tune the output circuit of a combined oscillator multiplier to a very much higher frequency than that of the grid circuit?

m. What differences exist in the requirements for a class C power amplifier used for f-m as compared to one used for a-m?

n. What is the principal disadvantage of the grounded-grid amplifier?

o. Why are push-pull amplifiers useful at high frequencies?

p. What is significant about the input impedance of a grounded-grid push-pull amplifier?

q. What are the major disadvantages of capacitive coupling as compared to link coupling?

r. What is necessary in respect to the power transfer between driver and final amplifier?

s. What are the functions of the power-amplifier output coupling network? Which of these features is determined by the operating Q of the circuit?

t. What are the relative advantages and disadvantages of shunt feed?

u. What is the function of a split-stator capacitor in a single-ended neutralized triode power amplifier?

v. How does the type of feed affect the d-c voltage appearing across the tank capacitor?

w. What is the major disadvantage of the pi-network as an output coupling circuit?

x. How can short antennas be coupled directly to the final amplifier with good efficiency?

y. What are the major operating indications that a final amplifier is oscillating parasitically?

z. How are parasitic oscillations suppressed?

aa. Why are certain combinations of input and output circuits avoided?

ab. What is the difference between a d-c control system and a mechanical control system?

ac. What are the relative disadvantages of a d-c control system?

ad. Compare the response speeds of a motor-control and a d-c control system. Which is more suitable for long-time changes?

ae. What is a receiver-transmitter frequency interlock system?

af. What device is most frequently used to compare the frequencies of two waves?

ag. What is the major disadvantage of the double-tuned discriminator?

ah. Why are discriminators used at a frequency that is low compared to the carrier frequency?

ai. What is the advantage of the phase discriminator as compared to the double-tuned discriminator?

aj. How does the modified phase discriminator differ from the standard phase discriminator?

ak. What is the advantage of a pulse control system as compared to a simple discriminator d-c control system?

al. What is the function of the modified balanced modulator in the pulse control system?

am. What is the function of the limiter tube in the pulse system?

an. Why is the trigger circuit used in conjunction with the pulse discriminator?

ao. How is the output of the pulse discriminator applied to the control of the master ocillator?

ap. Why is the motor control system restricted to large fixed installations?

aq. What are the three major parts of the motor control system?

ar. What is the purpose of using frequency dividers?

as. What is the difference between a free-running and a triggered divider?

at. What is the major disadvantage of the synchronized oscillator as a frequency divider?

au. How is fractional division obtained in a regenerative modulator?

CHAPTER 5

F-M RECEIVERS

Section I. INTRODUCTION

52. General

The operation and general circuitry of superheterodyne receivers for f-m communication form the subject matter of this chapter. The circuits used in the various operating parts of the superheterodyne receiver are discussed and comparisons are made between a-m and f-m circuits.

53. Superheterodyne

The separate stages required for the operation of an f-m superheterodyne are shown in the block diagram of figure 107. The r-f signal voltage coming from the antenna is amplified by the r-f amplifier stage and combined with the local oscillator voltage in the converter or mixer. The new frequency produced by this combination is the difference frequency or i-f (intermediate frequency) of the two signals, and is amplified by one or more i-f amplifiers. The greatly amplified i-f signal then is applied to an f-m detector which produces an audio-voltage output from the frequency variations of the i-f signal. This is applied to the audio amplifier, and the output of the amplifier is converted to the original sound by a loudspeaker or earphones.

54. Receiver Rating

a. Sensitivity and Selectivity. The sensitivity of a receiver is determined by the minimum signal voltage (in microvolts) at the input that is required to produce a specific output. The

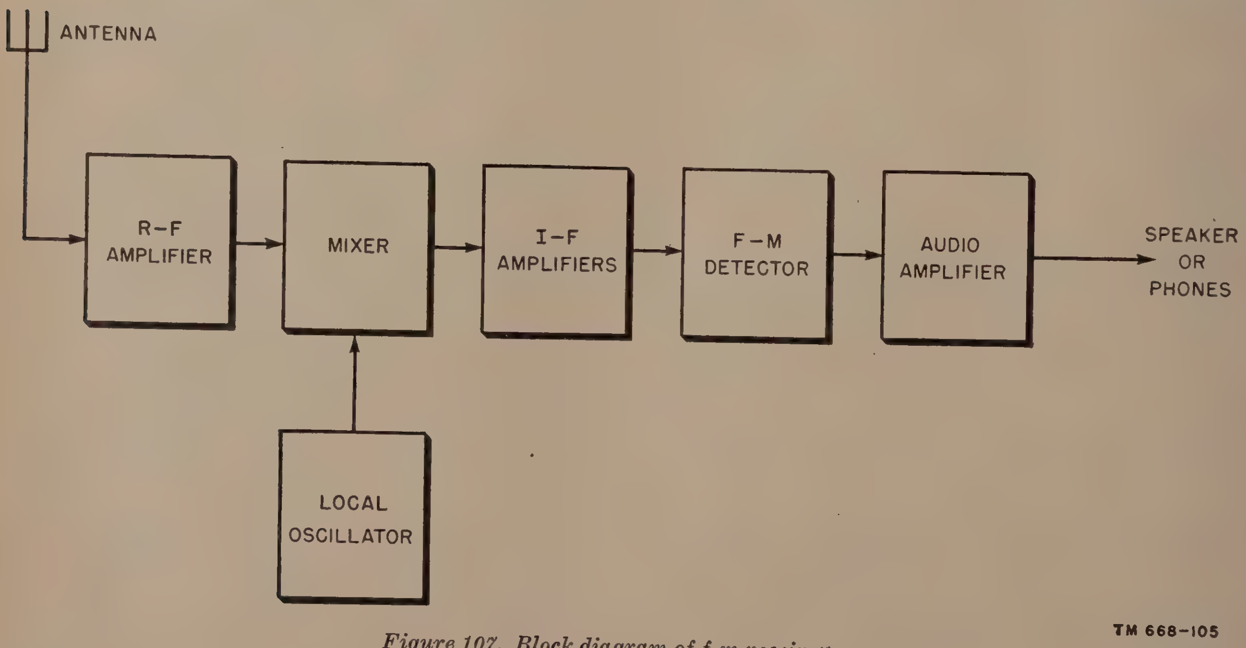


Figure 107. Block diagram of f-m receiver.

receiver producing the specified audio output with the smallest value of signal input is the most sensitive. The selectivity of a receiver is defined as the degree to which it is capable of discriminating against signals whose frequencies are other than that of the desired carrier. The selectivity for adjacent channels is determined largely by the characteristics of the i-f amplifier. Discrimination against image carriers depends on the circuit design of the r-f amplifier and the mixer, or converter.

b. A-M Response. Noise originating in actual defects, such as poor contacts or faulty parts, as well as the noise produced in the tubes, limits the maximum sensitivity of the receiver. Every f-m receiver, no matter how perfect, has some undesirable response to a-m signals, and these variations must be removed by limiting devices. The degree to which the receiver discriminates against amplitude variations in the received signal caused by fading or noise is a measure of the merit of the receiver.

55. Double Superheterodynes

Most f-m equipment operates in the v-h-f band, and at these frequencies it is not always possible to obtain optimum performance with the standard superheterodyne circuit. When good adjacent-channel selectivity is necessary, a low i-f is desirable; this, however, lowers the receiver rejection of image signals. Similarly, if good rejection of image frequencies is required, a high i-f should be used, but this is not compatible with good adjacent-channel rejection. These difficulties can be overcome by combining the advantages of high and low i-f amplification. The r-f signal is mixed with a local oscillator to produce a *high* i-f in the conventional manner. A *second* local oscillator voltage then mixes with the high i-f signal to produce a *low* i-f, as shown in figure 108. The double superheterodyne or double-conversion superheterodyne requires frequency conversion in two separate mixer or converter stages. To meet stringent military performance requirements, double superheterodynes commonly are used.

56. Auxiliary Circuits

Apart from the operational circuits, there are auxiliary circuits that provide automatic frequency control. To prevent operator fatigue, circuits are incorporated which silence, or squelch the noise that appears in the output of the receiver when no carrier is present. The special audio requirements of carrier telephone equipment also must be analyzed, since an important class of f-m equipment exists as a link between units using this type of telephone service.

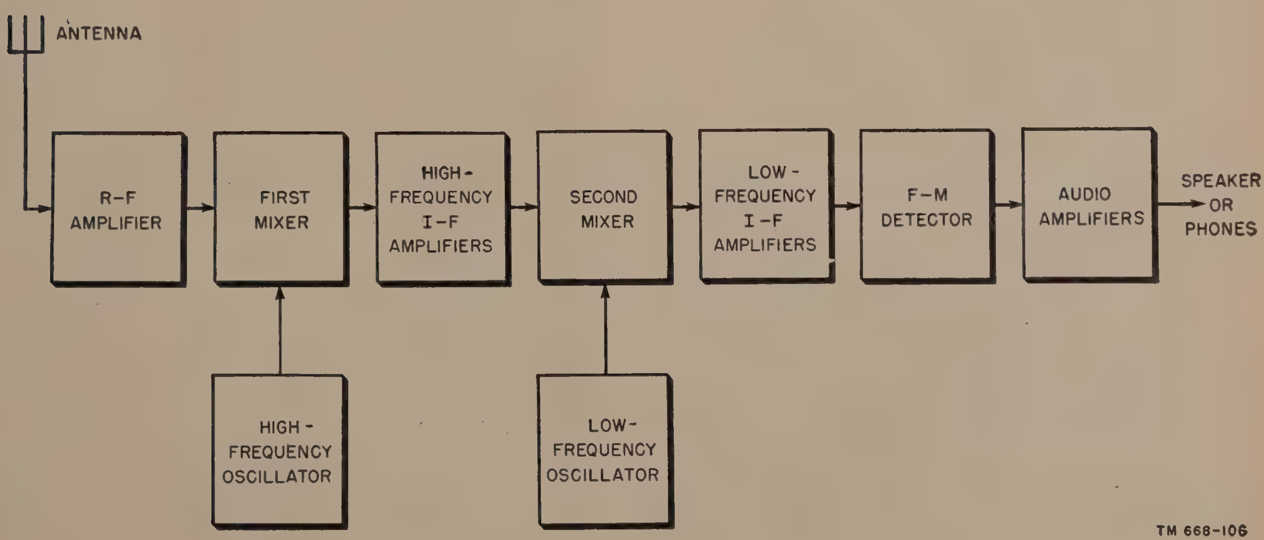


Figure 108. Block diagram of double-conversion f-m receiver.

57. Basic Functions

a. The basic functions of the r-f amplifier are to increase the signal-to-noise ratio, to provide adequate image-frequency rejection, and to suppress local-oscillator radiations. These functions determine the design of the many different r-f circuits that have been developed, and each r-f amplifier exhibits varying degrees of performance as regards these factors. Considerations of tuning range complexity and stability also influence the choice of a particular circuit.

b. Since the r-f amplifier receives a signal at the lowest level of any stage in the receiver, any noise or other disturbance introduced in this stage has a proportionately greater effect. The performance of the receiver in respect to weak signals depends on the performance of the r-f amplifier, or the signal-to-noise ratio of its output. The r-f amplifier also must reject images and other unwanted frequencies that can reach the frequency converter and produce spurious responses. The efficiency of rejection depends largely on the design of the tuned circuits used. The local oscillator used in superheterodynes causes considerable interference if its signal is allowed to travel back to the antenna. Therefore, the r-f amplifier must be designed to block the signal from the local oscillator and prevent it from radiating. This is important when several receivers are operated close together on many different frequencies.

58. Sources of Noise

a. Fluctuation Noise.

(1) In any conductor, the movement of the electrons which constitute the current flow through the conductor is erratic and random. This random motion of electrons in a conductor or in the electron stream of a vacuum tube gives rise to small fluctuations in the voltage developed across the resistance of the conductor. The random variation of voltage is called fluctuation noise.

(2) No definite waveform or frequency can be associated with the fluctuation noise. However, the average amplitudes at all frequencies (noise spectrum) usually can be specified. Since the noise is distributed uniformly in respect to frequency, the amount of average noise voltage depends on the bandwidth. The fluctuation noise present determines the minimum signal to which a receiver will respond. The signal voltage must be sufficiently large to prevent the noise voltage from overriding the modulation on the carrier.

b. Thermal Noise.

(1) The molecules of any physical substance are always in violent motion, the average rate of this motion being perceived as temperature. Motion of the electrons caused by heat produces *thermal noise* across the terminals of any conductor containing resistance to the electron motion. From Ohm's law, the greater the value of the resistance in ohms, the more voltage developed across the resistor. Therefore, as the motion of the electrons in the conductor increases with temperature, the amplitude of the thermal noise in the output increases. The effect of the thermal noise, which is a type of fluctuation noise, depends on the bandwidth of the circuit which develops the noise.

(2) Figure 109 gives the amplitude of the average noise voltage usually encountered in the input circuits of f-m receivers. The temperature is assumed to be normal room temperature, or 68° Fahrenheit. The bandwidth of the input circuit can be estimated from a graph of its response characteristic. For example, it is common for an f-m receiver to have a bandwidth of 200 kc and an input circuit resistance of 300 ohms. As indicated in the chart, a noise voltage of about 1.2 microvolts results. This receiver cannot produce a useful output with incoming signals under 1.2 microvolts because of the threshold effect, and a higher value of input is needed.

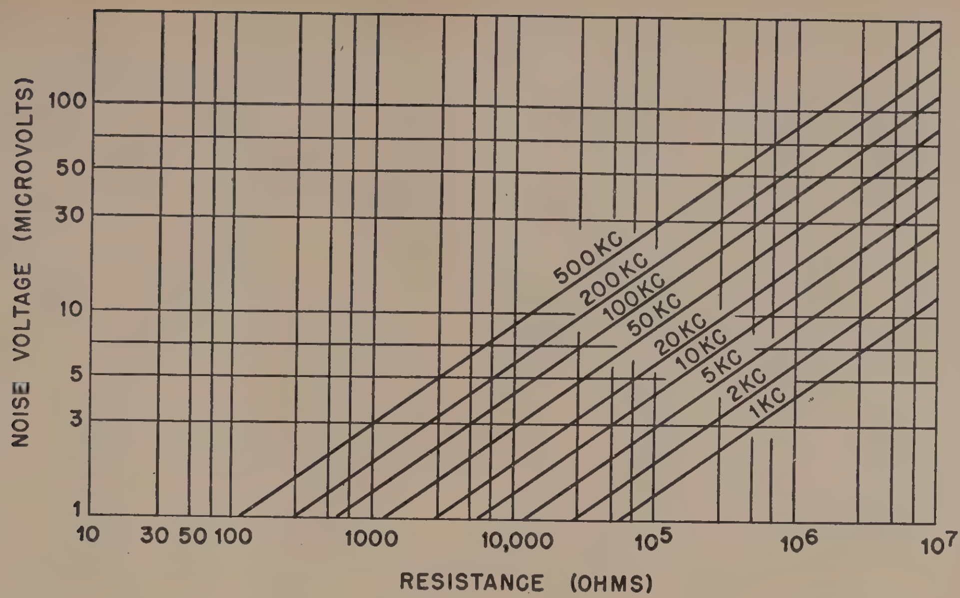


Figure 109. Noise voltage for different bandwidths and impedances.

c. Tube Noise. Fluctuation noise in vacuum tubes can be divided into three classes—shot noise, partition noise, and induced grid noise. Shot noise or shot effect occurs because the emission of electrons from the cathode is not uniform. Since the electrons leave the cathode at random instants of time and with random variations in velocity, their arrival at the plate is not uniform. Therefore, these variations in current flow represent a noise current and will produce a noise voltage when current flows through the load resistance. Partition noise is caused by the irregularities in the distribution of current in various positive electrodes in the tube. This noise is present only in pentodes, tetrodes, and other multigrid tubes. Variation in the number of electrons passing a control grid at high frequencies will induce large noise currents. This induced grid noise can be added to the plate current noise as though they were independent. The noise produced by a tube varies with changes in the potentials on its elements. A factor that strongly affects the amount of noise produced is the space charge in the tube. The noisiest condition is produced with temperature saturation; that is, when all of the available electrons are drawn to the plate and there is no space charge at all. It is the presence of space charge that makes amplification possible, and, fortunately, space charge also reduces the noise caused by random fluctuation in the electron stream.

d. Equivalent Noise Resistance.

(1) To permit direct comparison of the noise performance of standard tubes, a factor has been worked out that is known as the equivalent noise resistance of the tube. The value is that of a resistor that will produce the same amount of thermal agitation noise as that produced by the tube from all causes. The following chart indicates the range of values of this resistance for particular tube types.

Tubes	Ohms
Triodes	200 to 3,000
Sharp cut-off pentodes	700 to 7,000
Remote cut-off pentodes	2,400 to 14,000
Hexodes and heptodes	190,000 to 300,000

(2) The greater the number of positive grids, the greater the amount of noise produced by the tube, and the higher the equivalent noise resistance. The basic cause for this is the larger number of current divisions with increased partition noise. The noise produced by this equivalent resistance must be added to that introduced in the input

circuit of the tube to evaluate the performance of the r-f amplifier at low signal levels. In the previous example, the input circuit noise was 1.2 microvolts. Assume that this is applied to a pentode amplifier tube with an equivalent noise resistance of 3,000 ohms. From figure 109, the noise voltage corresponding to this resistance at the assumed bandwidth of 200 kc is 3.2 μv (microvolts). This is added to the input noise for an over-all noise of 4.4 μv.

59. Noise and R-F Amplifiers

a. General. Amplifiers do not discriminate between signal and noise within their bandwidth, but amplify them equally. For example, assume that the voltage amplification of the tube with its associated circuits in the previous example is 10 and that the tube delivers its output voltage into a similar amplifier. What will be the over-all noise performance of the system? With a noise voltage of 4.4 μv, a signal of 10 μv is applied at the input of the amplifier. Because the noise is of random phase it will be sometimes in phase with the applied signal and sometimes out of phase with it. Therefore, the signal in the output of the amplifier will be a combination of signal and noise which varies from the amplified sum to the amplified difference. The over-all output is the quadrature sum of the signal and the noise voltages, multiplied by the stage-amplification, or $\sqrt{10^2 + 4.4^2} \times 10 = 10.7 \times 10 = 107\ \mu\text{v}$. Since the second stage of amplification was assumed to be identical with the first, it adds 3.2 μv of noise to the applied signal of 107 μv. The output of the second stage is

$$(\sqrt{107^2 + 3.2^2})\ (10) = (107.1)\ (10) = 1070\ \mu\text{v}$$

The contribution to the total noise output of the 3.2 μv of noise added by the second stage is obviously very small. Therefore, the noise performance of the two stages together is essentially that of the first stage alone. If the second stage is the mixer in a superheterodyne circuit, it follows that the noise performance of the receiver is determined almost entirely by the first r-f amplifier. In general, if the r-f gain exceeds 5, the effect of the second stage is slight; if it exceeds 10, the effect is negligible.

b. Noise Figure.

(1) The noise figure expresses the relative merit of a receiver in comparison with a so-called perfect receiver. The perfect receiver is one which adds no noise to that produced by the antenna resistance and has a noise figure of 0 db. The quantity normally is expressed as a power ratio converted to decibels, and the smaller the noise figure, the better the receiver. Generally, the higher the frequency, the harder it is to obtain a good noise figure. Best attainable values range from 10 to 12 db at centimeter wavelengths, 6 db in the upper portion of the v-h-f range, where most f-m equipment is used, and below 3 db for frequencies under 30 mc. A perfect receiver has a noise figure of 0 db.

(2) The noise figure does not depend on the bandwidth of the receiver under test. A receiver that has a bandwidth of 200 kc has no greater sensitivity than one with a bandwidth of 15 kc if it has the same noise figure. However, because of the greater amount of noise contained in a wider bandwidth, it is harder in practice to obtain as low a noise figure for a wide-band receiver.

c. Tube Types and Noise Figures.

(1) The noise figure of an r-f amplifier is affected by many conflicting variables, and compromises must be reached. The tube is the ultimate factor which dedetermines the minimum noise figure that can be achieved, the transconductance and the input conductance being the principal factors in determining the noise figure. For good noise figure, it is desirable that the r-f amplifier have high gain so that the noise added by the following stages will be negligible. The higher the transconductance, the more gain from the stage. The input conductance of a tube is defined as the equivalent impedance that is seen looking into the grid. Effectively, it is composed of a resistance and a capacitance in parallel, placed across the tuned input circuit, and it

decreases as the operating frequency increases. Doubling the frequency decreases the input resistance by a factor of almost four; tripling it decreases the resistance by a factor of nine, and so on.

(2) It is common to use tuned transformers in the input circuits of r-f amplifiers to provide a voltage step-up from the low-impedance transmission line to the input impedance of the grid. Since the input impedance of the grid varies with frequency, the amount of voltage step-up that can be obtained decreases as the input conductance increases. Since no noise is generated in the input transformer, and the amount of noise generated by the tube is constant, the higher the input voltage applied to the tube, the lower the effect of tube noise, and the better the noise figure. It is desirable to use tubes that have a low input conductance. This permits a high voltage step-up in the input transformer with consequent overriding of tube noise. The input conductance will vary as the square of the frequency and has been found to depend on the input capacitance, the transconductance, and the inductance of the cathode load. For the lowest input conductance, all of these factors should be minimized. However, to obtain gain from the amplifier, there must be a reasonable amount of transconductance, and the choice of these values is necessarily a compromise. Since the noise produced within the tube is caused largely by irregularities in the electron stream, this stream must have great uniformity for low noise figures. The relative performance of various tube types in common use is listed in the table of equivalent noise resistances. Ratings in terms of input conductance generally are not available for all types.

60. R-F Amplifier Circuits

a. General. The r-f amplifiers of f-m receiving sets operate in the class A region of the operating curve and do not draw grid current. The several possible connections for a tube used as an r-f amplifier depend on which element is common to ground. For the triode, the possibilities are grounded cathode, which is the conventional connection, grounded grid, and grounded plate (cathode follower). The pentode and tetrode also follow these classifications, since the screens and other grids usually are at ground potential for radio frequencies.

b. Input Circuit Operation.

(1) The input circuit of most r-f amplifiers consists of an r-f transformer with an untuned primary that is connected to the antenna. The secondary of this transformer is considered to be a tuned series-resonant circuit that is adjusted to resonate with the frequency of the desired signal. Since the transmission line from the antenna is connected to the primary and generally has a low impedance (50 to 600 ohms), and the grid-to-cathode impedance of the tube is much higher, there is an impedance mismatch with serious loss of signal. The r-f transformer provides the necessary impedance match between the antenna and the grid of the tube. Actually, there is a slight mismatch for optimum noise figure.

(2) The transmission line used to transfer the signal from the antenna to the primary of the r-f transformer may be either coaxial-line or parallel-wire transmission line. When the coaxial line is used, the outer conductor is at the same potential as the circuit ground and an unbalanced input is required at the receiver. If parallel-wire transmission line is used, the voltages on each conductor are equal and opposite to each other and 90° out of phase with the ground potential, and a balanced input is required at the receiver. Both input circuits provide a certain amount of discrimination against noise and other spurious frequencies picked up on the transmission line. The outside of the coaxial line can act as an antenna and

receive signals on the shield conductor. However, this outer conductor is effectively grounded at the input circuit and therefore produces no voltage across the grid-to-cathode circuit of the tube. Balanced input circuits suppress noise picked up simultaneously on both conductors of the line. The noise voltage is in phase on both sides of the line. This means that the voltage in respect to ground is 180° out of phase. The balanced input circuit responds only to currents that are out of phase on the line and 90° out of phase to ground; therefore, the noise voltage does not get to the grid of the tube.

(3) The output of the series-resonant circuit of the secondary decreases for frequencies on either side of resonance. Therefore, the over-all response of the amplifier falls off as the frequency increases or decreases in respect to the operating frequency. This produces the characteristic bell-shaped response of voltage versus frequency that is referred to as the selectivity curve. The ability to reject signals that are far removed from the operating frequency depends largely on the sharpness of this curve. In general, the sharpness of resonance of a tuned circuit depends on the Q of the inductor and the extent to which the input conductance of the r-f amplifier tube loads the input circuit. However, it usually is possible to manufacture for high-frequency work inductors whose Q is sufficiently great that the resistive component of the input conductance has a greater effect on the selectivity than does the resistive component of the coil impedance.

(4) Although it is possible to obtain the maximum amount of selectivity with a given tube, it often is necessary to compromise, since the receiver does not operate on a fixed frequency, but is tuned over a considerable range. This is accomplished by using a coil with more than the optimum amount of inductance. The various tuned circuits in the r-f mixer and oscillator stages must be correctly tuned for any operating frequency in the desired range. This means that the tuned circuits of the local oscillator and the r-f amplifier must *track* correctly over the entire range.

(5) For receivers that cover a limited band of frequencies, it often is desirable to avoid the complications of tracked circuits. This can be accomplished by keeping the r-f amplifier tuned to a fixed frequency and varying the frequency of the local oscillator. The selectivity of the input circuit is made low enough to cover the entire band in question and is called a *broad-band input circuit.* When the primary and the secondary of a transformer are moved close to each other, the coupling is increased beyond the critical value, and the response curve begins to broaden out. Instead of the familiar bell curve, a broad, flat-topped curve appears and the circuit is said to be overcoupled. As the coils are moved closer together, they become extremely overcoupled, the center of the curve begins to drop, and the two peaks appear still further removed from the center frequency. Such circuits are called overcoupled input circuits, and curves for values of coupling are shown in B of figure 110. A simple resonant circuit cannot provide both impedance transformation and wide bandwidth at the same time, and overcoupled input circuits with double and even triple tuning must be used. The circuit shown in A is overcoupled and a capacitor is placed in series with the primary and adjusted to resonance at the operating frequency. This provides the bandwidth necessary for f-m reception. More complicated versions, which produce three peaks in the response curve, can be obtained with three tuned circuits in the input. In f-m

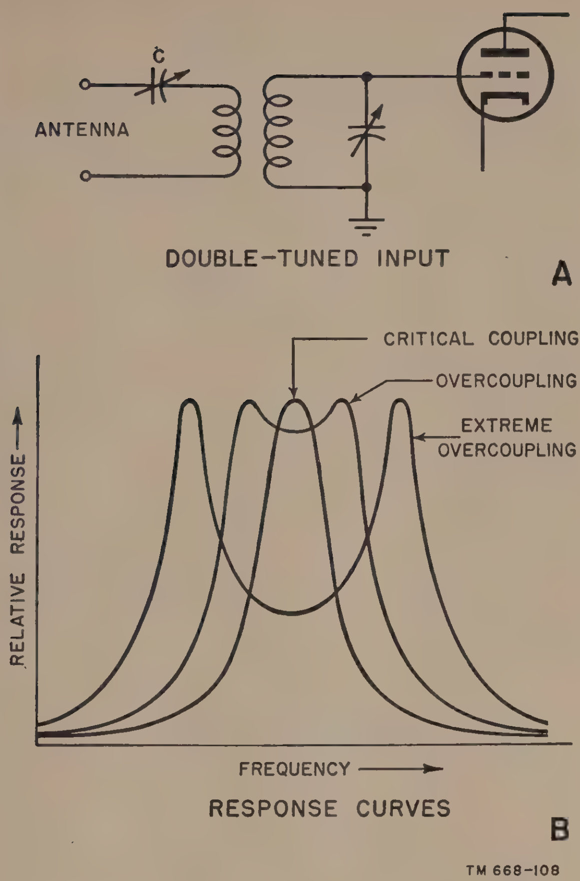


Figure 110. Coupled input circuit for r-f amplifiers.

equipment, the single- or double-tuned types are the most frequently encountered.

(6) At very-high frequencies, it is difficult to get the transformer type of input circuit to operate properly because of leakage and stray capacity. Therefore, the antenna or transmission line often is tapped directly on the coil. This connection (A of fig. 111) is suited especially to coaxial inputs. The operation is essentially that of an autotransformer, the coupling being between the two sections of the coil instead of through a separate primary as in a conventional transformer arrangement. Since the bottom of the coil is at ground potential and the top at grid potential, the input impedance increases as the tap is moved up the coil, and any value of line impedance can be matched. Usually, a small fixed capacitor is inserted in series with the tap lead to prevent stray d-c voltages that may appear on the antenna or transmission line from reaching the grid.

(7) If a large variety of antennas and transmission lines is to be used with the amplifier, it is necessary to provide an input circuit that can match a wide range of impedance. It is possible to construct a tuned transformer with a primary that can be moved in respect to the secondary. As the coupling varies, the effective input impedance also varies. Since this arrangement can be complicated mechanically, however, a variation of the tapped input transformer is used instead. To avoid the complication of a variable tap on the coil, two variable capacitors in series are used as tuning elements with the line connected to their junction, as in B. The capacitors act as a voltage divider. If the lower one has a much higher capacitance than the upper one, it has less reactance at the operating frequency. Therefore, the impedance across it is low and can be adjusted to match a wide variety of transmission lines. A variation of this circuit is shown in the pi-network of C, where the rotors of the two capacitors are grounded, with the input at the junction of the first capacitor and coil, and the output taken from the opposite junction. The pi-network passes all frequencies below the resonant frequency with only slight attenuation. This can be disadvantageous if the equipment is operated near powerful low-frequency transmitters. Input circuits for operation with balanced lines are shown in D, E, and F. Operation of the balanced circuits is similar to that of their unbalanced counterparts.

c. Grounded-Cathode Amplifier. The grounded-cathode r-f amplifier in A of figure 112, is the most familiar of all types. The input may be applied with any one of the many dif-

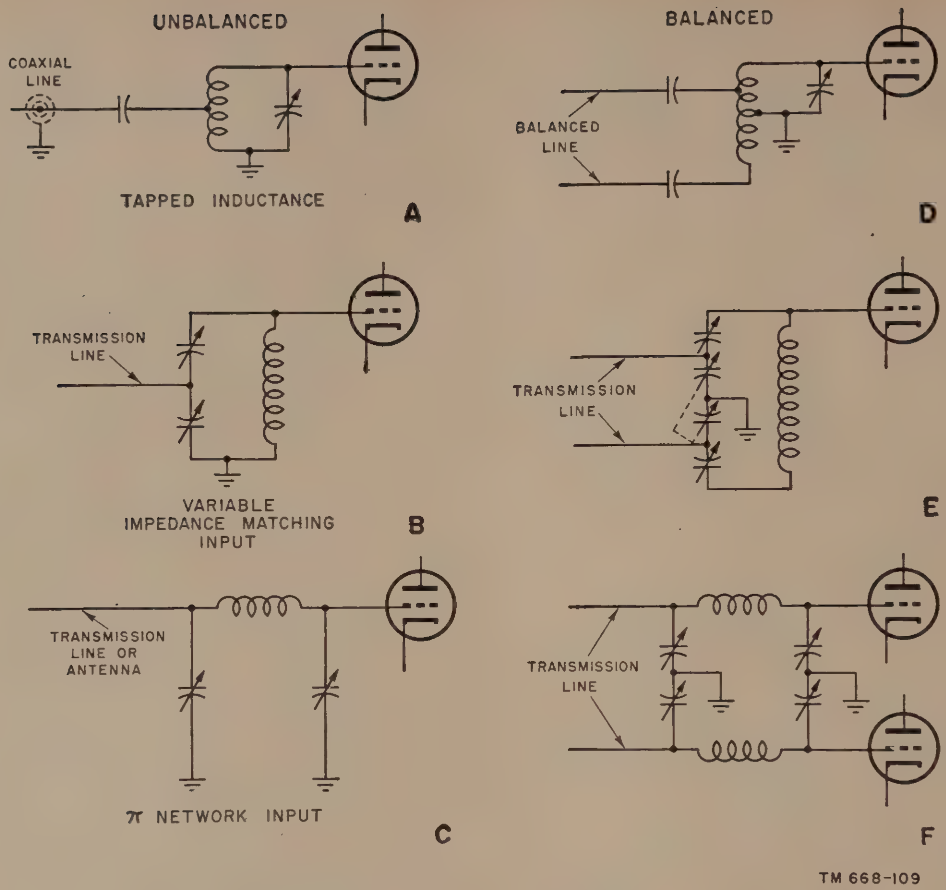


Figure 111. Miscellaneous input circuits for r-f amplifiers.

ferent input circuits to the grid, and the output is taken from a resonant load circuit in series with the plate. If a triode is used in this way, it must be neutralized to prevent tuned-plate tuned-grid oscillation. This neutralization seldom is successful over a wide frequency range, and the circuit generally is used only with pentodes. The noise performance of this circuit is good when triodes are used, but, with pentodes which have inherently higher noise figures, the arrangement is poor compared with that of other r-f amplifiers.

d. Grounded-Grid Amplifier. The grounded-grid circuit, shown in B of figure 112, permits the use of the triode with its lower noise figure, and does not require neutralization. However, the voltage gain of the amplifier is not as great as that of the grounded-cathode circuit because the input impedance is very low. The tuned circuit has little voltage step-up to overcome tube noise and the over-all noise performance suffers. The low-impedance input circuit permits the attainment of wide bandwidth and a reasonable noise figure without sacrificing too much voltage gain in the input circuit. The gain of the grounded-grid amplifier may not be great enough to override the noise produced by some converter tubes; therefore, it is common practice to find two grounded-grid r-f amplifiers in cascade. The added complications arising from this necessity and the need for special tubes limit its use. The tubes themselves must have very low effective plate-to-cathode capacitance if the shielding effect of the grounded grid is to be realized.

e. Cathode Follower. The amplifier shown in C has the input applied between the grid and ground, the plate is grounded for r-f, and the output circuit is in series with the cathode. The voltage gain of the stage is always less than unity, because the voltage variation at the cathode in an amplifier always is less than that

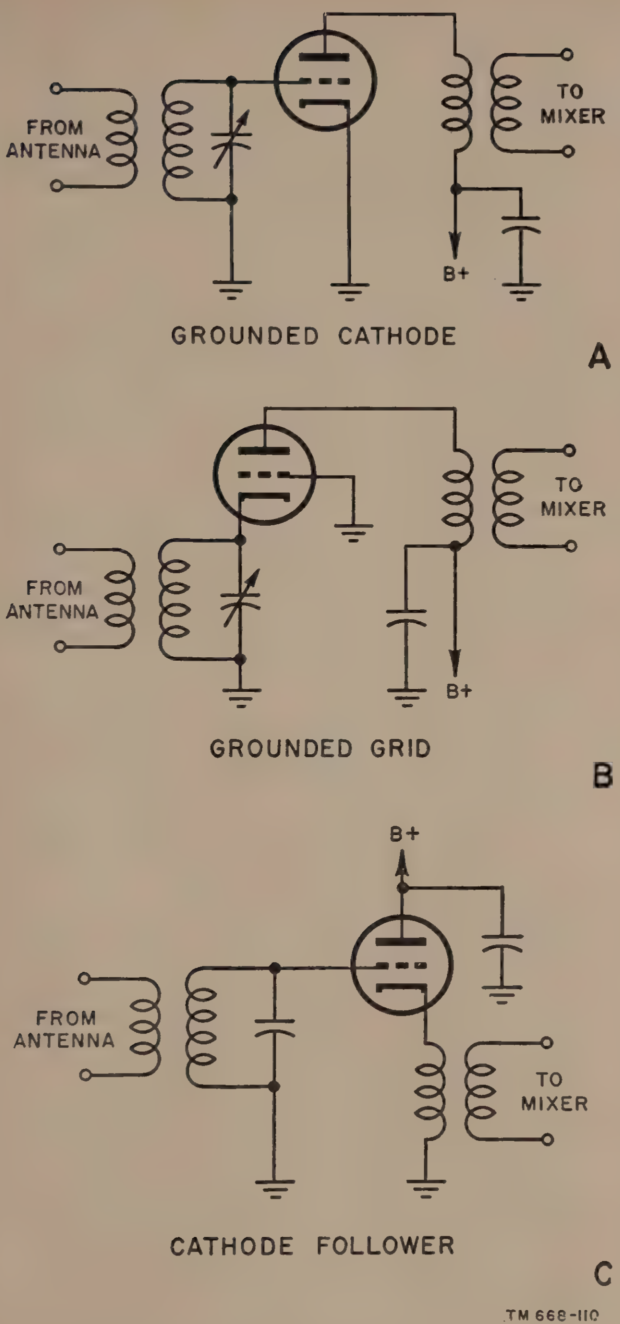


Figure 112. Single-tube r-f amplifier circuits.

at the grid. The output impedance of this amplifier is low because the cathode voltage is low and the current high. The input impedance is many times greater than the conventional grounded-cathode amplifier. Therefore, the input circuit can have a higher impedance and a higher voltage gain. The loss in the tube output, however, cancels out this advantage, giving a net result not very different from that of the conventional grounded-cathode amplifier. The cathode follower can be used with a triode without fear of oscillation. The low-impedance output has to be stepped up with an additional tuned transformer to match the input of the converter tube. Therefore, there is some danger of instability in coupling between the high-impedance transformers in the input and output circuits.

f. Single-Tube Amplifiers. Except for some differences in stability, input bandwith, and the like, there is little difference in the sensitivity that can be obtained with the three types of amplifiers. The noise figure of the tube is essentially independent of the manner of circuit connection. Where the high gain of a pentode stage is needed, the grounded-cathode circuit is used. Where a good noise figure and broad input bandwidth or a good match to the transmission line is desired, the grounded-grid circuit serves well. If high input selectivity is desired, the cathode follower, which loads the input circuit the least, can be chosen.

61. Two-Tube R-F Amplifier Circuits

a. General. The grounded-cathode, grounded-grid, and cathode-follower amplifiers may be used together or in push-pull. The push-pull amplifier permits a low-impedance, balanced-input circuit and can be used at very-high frequencies. Circuit diagrams are shown in figure 113. When cascade amplifiers are used, it is possible to have nine different circuit arrangements, since the output of any one of the three amplifiers may be connected in three different ways. The circuits most frequently used are the double grounded-grid, the cathode follower into grounded-grid, the grounded-cathode into grounded-grid, and the double grounded-cathode circuits.

b. Push-Pull Amplifiers.

(1) The push-pull circuit has a balanced input and output, and therefore is used widely with a balanced line. Since input voltage for a push-pull circuit must be twice that for an unbalanced amplifier, and since the input impedance is four times that for a conventional circuit, the over-all re-

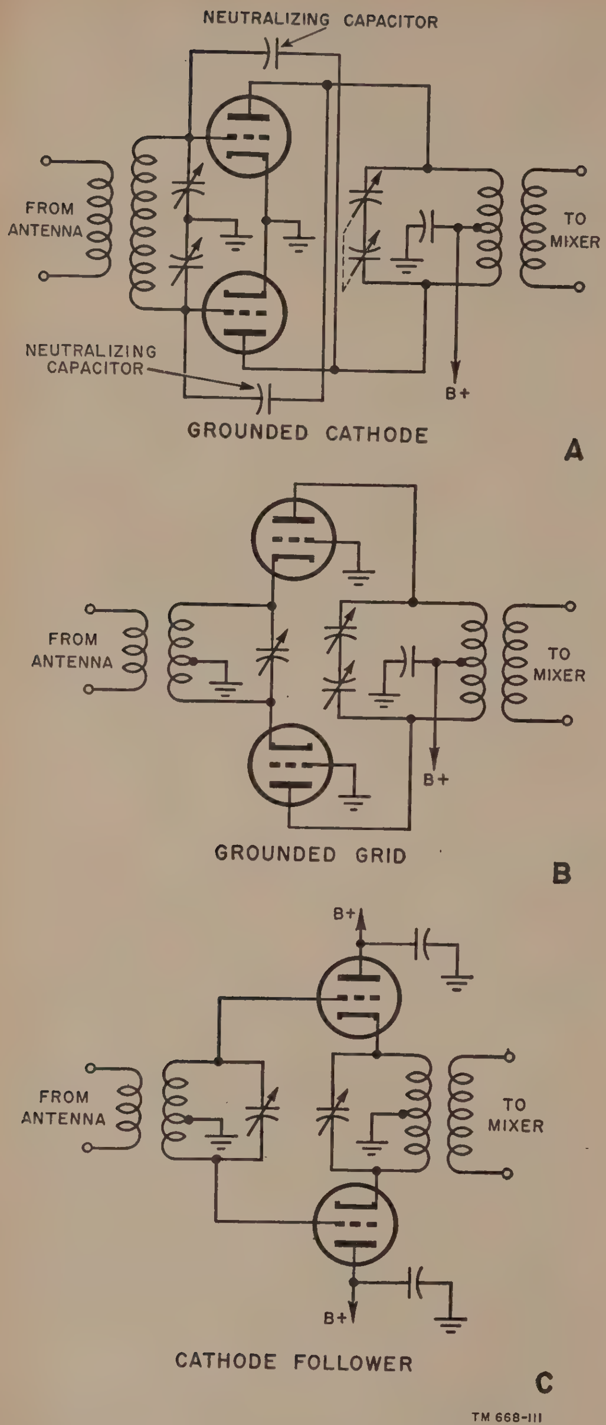


Figure 113. Push-pull r-f amplifiers.

quirement of impedance and voltage results in an input circuit with a higher ratio of inductance to capacitance. This becomes significant at very-high frequencies, where the internal tube capacitance can be an appreciable part of the circuit capacitance. The tube capacitance is halved effectively in the push-pull circuit, because the grid-cathode capacitance of the two tubes is in series across the input circuit. In the grounded-cathode circuit using triodes, it is necessary to use a cross-neutralization circuit. The over-all result is a circuit with a voltage gain that is much the same as that for a single neutralized triode, for the input voltage must be doubled even though the outputs of the two tubes add to each other.

(2) This implies that the noise produced by each tube also adds in the output. However, the voltage step-up in the input circuit is greater, and the over-all noise figure for the push-pull grounded-cathode circuit therefore is the same as that for each single-ended tube. The major advantage of the push-pull circuit, apart from its balance, is comparative ease of neutralization. Tubes such as the JAN types 6J6 and 12AT7 are specifically designed for these circuits.

(3) The push-pull grounded-grid amplifier, in B, permits the use of a balanced input circuit of very low impedance. This means that a broad bandwidth can be obtained easily. The voltage gain of the amplifier is the same as that for a single grounded-grid stage, and the noise figure is unchanged. This circuit can be used at very-high frequencies, where neutralization required for the grounded-cathode push-pull stage becomes too critical. The tubes used must have low plate-to-cathode capacitance, or the shielding effect of the grounded grid is lost. Moreover, where two triodes are incorporated in a single envelope, the cathode leads for each half must be brought out separately. JAN tube types 12AT7 and 6BQ7 sometimes are used in this way.

(4) Push-pull cathode followers are uncommon as r-f amplifiers because the

low-impedance output is not suitable for most mixer and converter tubes. However, when the mixer is a germanium or silicon crystal, it has a low-impedance input and the push-pull cathode follower can be used.

c. Two-Tube Cascade Amplifiers.

(1) Because of the deficiency in gain of a single grounded-grid stage, two stages generally are used in cascade, The output of the first is applied to the input of the second, as shown in A of figure 114. The over-all gain is the product of the individual stage gains. Because the gain of the first stage is seldom sufficient to permit it to overcome its own internally generated noise, the dual-cascade grounded-grid amplifier is not capable of as low a noise figure as a single neutralized triode. The arrangement does not require neutralization, and, because of the low impedance of the input circuits, it has a large usable bandwidth.

(2) The cathode follower driving a grounded-grid amplifier is referred to frequently as a cathode-coupled amplifier. In this connection, the tubes have a common cathode connection through the cathode resistor, as shown in B. The noise figure for this amplifier is good, although each stage contributes to the noise without enough added amplification to overcome it. Consequently, where low noise is necessary, this circuit is less desirable. The input impedance is high and the output impedance also is comparatively large. Only two tuned circuits are necessary, with the common cathode resistor acting as the interstage coupling. In addition, amplifier operation is stable without the necessity for neutralization. Where simplicity of circuit connection is required, the cathode-coupled amplifier combines fair performance with a minimum number of parts.

(3) The most important of the cascade amplifiers consists of a grounded-cathode triode driving a grounded-grid triode. This circuit usually is referred to as a driven grounded-grid circuit. From the standpoint of noise figure, it has the most favorable combination of characteristics. Several possible arrangements, shown in C, D, and E, permit this circuit to operate with stability. The amplifier is stable because of the low plate load of the input amplifier formed by the input of the grounded-grid second stage. The voltage gain of the first tube is negligible, and gain is obtained in the second stage with the noise figure of the first stage. If the amplification factor of each tube is large, the noise figure is that of the input triode alone. This circuit has all of the advantages of the triode grounded-cathode amplifier without any of the disadvantages, such as need of precise neutralization. However, the noise figure of the first stage can be improved further at higher frequencies if it is neutralized.

(4) The three diagrams, C, D, and E, indicate different means for accomplishing neutralization in the driven grounded-grid circuit. C shows an inductor and blocking capacitor connected in series between the grid and the plate of the first tube. The value of inductance is chosen to produce parallel resonance with the grid-plate capacitance at the operating frequency. A variant of this method of neutralization is shown in D. Here, an out-of-phase voltage from the output circuit is applied to the input circuit through a capacitor. This arrangement operates over a wide range of frequencies if the leads of the neutralizing capacitor are short. The circuit of E is similar to that of D except that the two sections of the tube are coupled directly from plate to cathode, and the plate current of the first tube flows through the second tube. This arrangement is the ultimate in simplicity and noise performance, since it uses very few parts. The

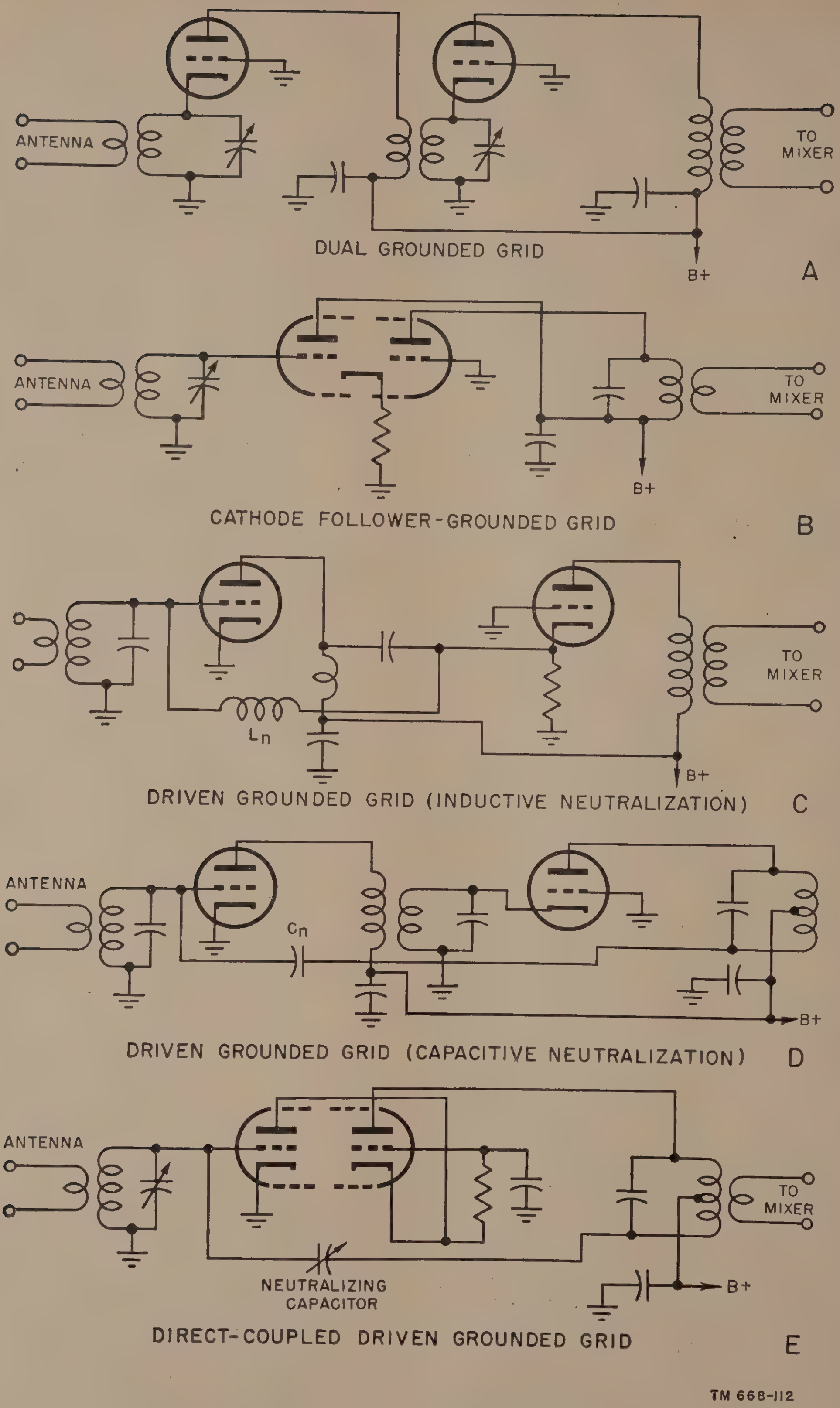


Figure 114. Cascade r-f amplifiers.

direct-coupled circuit combines the cut-off characteristics of both tubes. Therefore, variations in d-c grid voltage required to cut off the two tubes are greater than for a single tube alone. This reduces the susceptibility of the stage to overload from strong signals and reduces modulation between signals of different frequencies.

d. Automatic Volume Control for R-F Amplifiers.

(1) Automatic control of the gain in the first r-f amplifier of an f-m receiver is desirable because it tends to equalize the signal applied to the mixer or converter stage. In a-m receivers, avc (automatic volume control) is used to hold the over-all gain of the receiver to a constant level and to minimize fading and similar effects. In an f-m receiver, this is not necessary, because the f-m detector does not respond readily to fading in a signal. Therefore, the requirements placed on the avc circuit in an f-m receiver are not severe, and many f-m receivers do not contain avc.

(2) Since strong signals tend to overload the mixer stage, the f-m avc circuit applies a negative potential to the control grid of the first r-f amplifier. This negative potential increases with an increase in signal, reducing the gain of the stage. Although tubes with a remote cut-off characteristic give the best performance, they have poor noise characteristics. The direct-coupled dual-triode arrangement is a practical solution to this problem. Sharp cut-off tubes can be used, although smaller avc voltages must be applied to prevent cutting off the tube entirely on strong signals.

Section III. MIXERS AND CONVERTERS

62. General

a. Principles and Purposes of Frequency Conversion.

(1) The frequency converter in a superheterodyne receiver beats the signal from the r-f amplifier against the signal from the local oscillator to produce a signal at the intermediate frequency. Frequency conversion circuits are made up of two parts, the local oscillator and the mixer. The local-oscillator signal and the r-f signal amplitude-modulate the electron stream in the mixer. This action produces side bands equal to the sum and difference frequencies of the r-f and local-oscillator signals. The lower side band, which is the difference frequency, generally is used as the i-f. The local oscillator always is much stronger than the signal so that the percentage of modulation is low. This prevents the development of spurious side bands. If the oscillator signal contains harmonics, side bands also are produced between these and the signal. It is the function of the tuned input circuit to pass only the signal frequency, and the output circuit discriminates against all but the desired side band.

(2) Although separate local oscillators are used almost always at the higher frequencies, the functions of the mixer and the oscillator can be combined in one tube envelope. This tube is called a *converter* and usually is limited to intermediate frequencies below 15 mc in f-m receivers. In a double-conversion f-m receiver, however, the second mixer operates in the frequency range where a converter is practical.

b. Operation of Modulation Process.

(1) The low percentage of modulation required for frequency conversion can be produced in several ways. The method most frequently used depends on the *transfer* characteristic of a

tube or other circuit element. The transfer characteristic expresses the relationship between the signal applied to the input of a device and the signal obtained from its output. For example, a straight-line transfer characteristic of a resistive network is shown in A of figure 115. This is called a linear transfer characteristic and shows that the output voltage is directly proportional to the input voltage. A sine-wave input to this circuit results in the output shown in B.

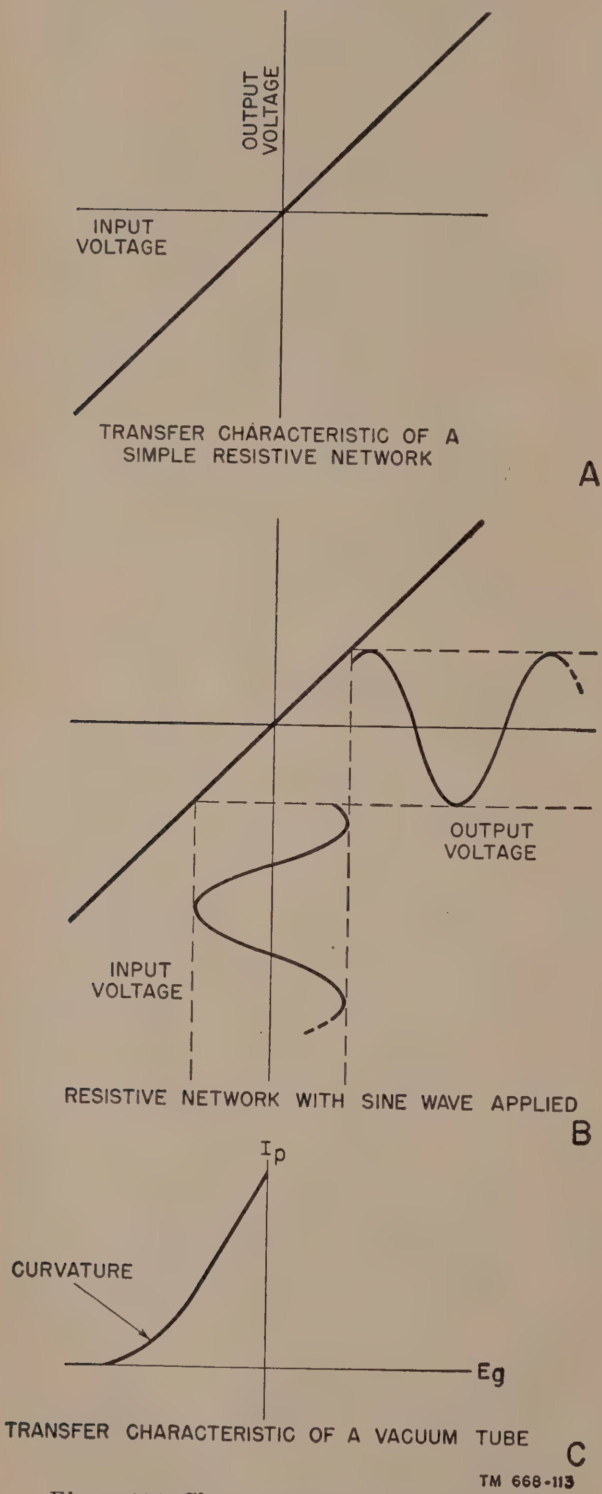


Figure 115. Characteristic response of resistive network.

(2) The transfer characteristic of a vacuum tube is not a straight line, since the relationship of e_g to i_p usually is curved at low values of plate current (C of fig. 115). Therefore, the vacuum tube is a *nonlinear* device. When the voltage on the grid of a vacuum tube becomes more negative and reaches the cut-off value, no current flows in the plate circuit. Consequently, for an entire range of voltages no current flows in the output circuit. Therefore, unlike the resistor, where current flows in proportion to any value of applied voltage, the vacuum tube is nonlinear, even if its transfer characteristic is perfectly straight. Other devices besides vacuum tubes are nonlinear, including silicon and germanium crystals.

(3) Any nonlinear device can be used as a mixer. In figure 116, two sine waves of different frequencies are applied to a nonlinear circuit. The output waveform is the familiar modulation envelope, which is equivalent to the carrier plus the side bands.

(4) There are two methods of producing the modulation envelope in a vacuum-tube mixer, and both utilize the electron stream. One method injects the r-f signal at the grid adjacent to the cathode and the local-oscillator signal at the same, or a following grid. The other method connects the oscillator to the grid adjacent to the cathode or to the cathode itself, and applies the r-f signal to an outer grid. Each method produces a variation of the electron stream by the oscillator, and a further variation by the r-f signal.

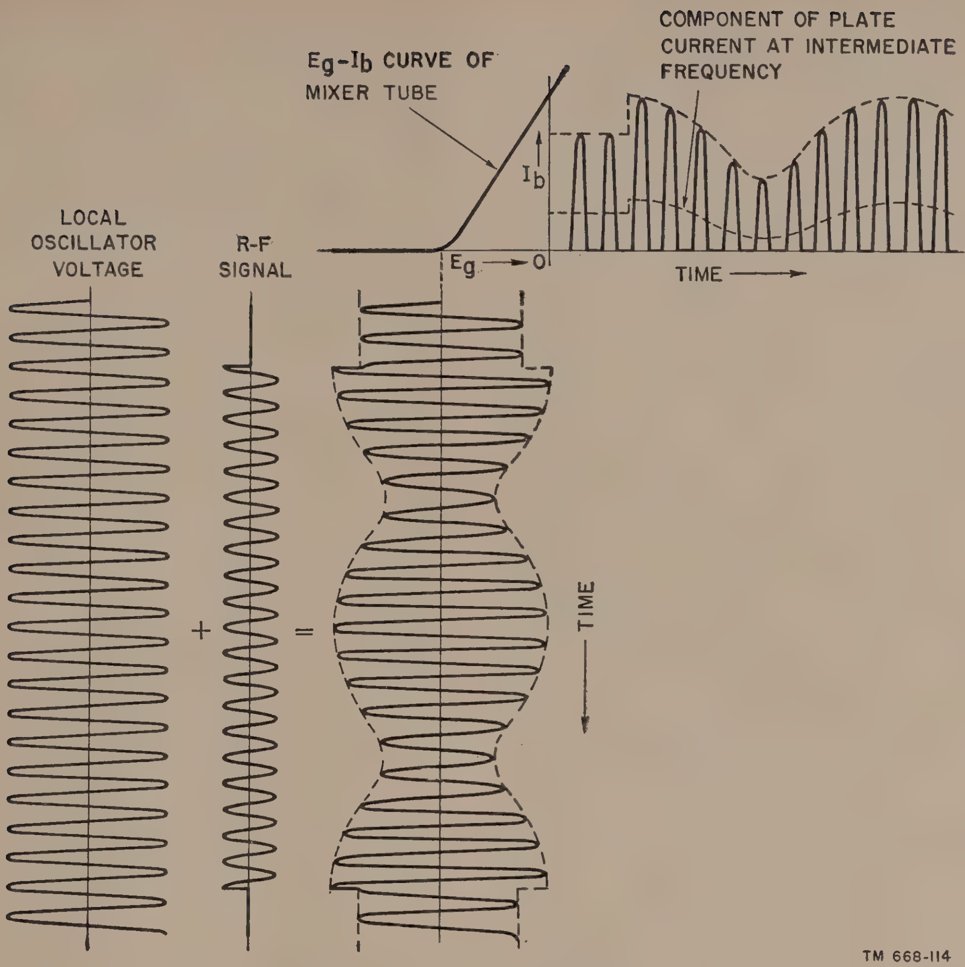


Figure 116. Mixer operation.

The electrons which reach the plate produce a current which has the shape of the modulation envelope. Effectively, these two methods do not differ, because of the nonlinear properties of an electron stream attracted to a plate.

c. Classification of Mixers. Mixers are classified according to the way in which the r-f and local-oscillator signals are applied. When both are applied to a single terminal, the mixer is called single-ended. If the r-f signal is applied to the control grid and the local-oscillator signal to an outer grid, the mixer is called *inner-grid* modulated. When the oscillator signal is injected before the r-f signal, the tube is said to be *outer-grid* modulated. It is desirable to inject the local-oscillator signal at a low-impedance point such as the cathode of the tube. This type of oscillator-signal connection is called a *cathode-injection* mixer.

d. Conversion Transconductance.

(1) The *conversion transconductance* of a mixer is defined as the ratio of the output current at the intermediate frequency, I_{if}, to the input signal, E_{rf}, at radio frequency with zero plate load impedance:

$$G_c = \frac{I_{if}}{E_{rf}}$$

Since frequency conversion is involved in a mixer, this must be indicated in the conversion transconductance. Strictly speaking, conversion transconductance is defined with small load resistance compared to plate resistance. As long as this condition is met,

changes in output are caused only by changes in load resistance. Maximum output at the intermediate frequency is obtained with cut-off bias on the control grid. Therefore, the operation of the mixer is similar to that of the plate detector. There is a very close similarity between conversion transconductance and the ordinary transconductance of an amplifier tube.

(2) The amplitude of the local-oscillator signal must be great enough to reach the part of the transfer characteristic with the steepest slope. As grid current begins to flow in the local-oscillator circuit, the input circuit of the mixer becomes loaded and therefore the local oscillator plus the r-f signal amplitude must be less than this amount. The mixer control grid must never be driven positive in respect to the cathode by local oscillator voltage. Bias for the mixer is obtained with a cathode resistor or, more frequently, with a grid leak. When a grid leak is used, the amplitude of the local-oscillator signal determines the d-c bias. The value of the a-c plate current which flows at the intermediate frequency, and hence the total output, depend chiefly on the slope of the transfer characteristic. The maximum conversion transconductance that can be reached is approximately 28 percent of the transconductance of the tube operating as an amplifier.

e. Conversion Gain.

(1) The output current of the mixer, as shown in figure 116, contains a strong component at the intermediate frequency. The tuned circuits between the output of the mixer and the i-f amplifier are parallel-resonant at the i-f frequency, and have a high impedance at the intermediate frequency and low impedance at all others. Therefore, the output current develops a large i-f voltage drop across this circuit. The conversion gain of the mixer is defined as the ratio of this i-f output voltage to the r-f signal input voltage:

$$A = \frac{E_{if}}{E_{rf}}$$

The amount of gain depends on the impedance of the tuned circuit. Therefore, it cannot be specified with a single constant for each tube type. A graph can be drawn, however, which shows this gain for different values of load impedance.

(2) The gain of the mixer depends on the conversion transconductance of the tube, since this determines the amount of current which flows through the tuned load circuit and the voltage developed across it for a given value of grid voltage. For high sensitivity, the mixer must have a high value of conversion transconductance. Similarly, a high value of load impedance also is desirable. The flow of current in the output circuit of the mixer is limited by the *a-c plate resistance* of the tube. This is the ratio of the a-c plate voltage to the a-c plate current at the intermediate frequency and is essentially in series with the tuned output circuit. When its value is more than five times that of the impedance of the output circuit itself, the current which flows is limited by the plate resistance alone. Therefore, the conversion gain does not depend on the plate resistance of the tube for low values of load impedance. As the load impedance is made larger, however, it begins to approach the value of the plate resistance, and the current flowing in the circuit decreases, with a corresponding decrease in the conversion transconductance. The load impedances which give the limiting value of gain usually are much higher than can be obtained with practical tuned circuits. Therefore, the maximum gain of the mixer stage as a whole is effectively determined by the maximum attainable impedance of the tuned plate load circuit.

63. Requirements for Mixers and Converters

a. Spurious Responses.

(1) Since a mixer is nothing more than a low-level modulator, the tuned input circuit can combine many signals with the fundamental of the local oscillator, or any of its harmonics, to produce an i-f output. If all but the desired signal are greatly attenuated before reaching the mixer input, and the oscillator is operating with low harmonic content, the response to spurious signals is minimized. This response depends on the selectivity of the input circuit of the mixer and of the preceding r-f stage. The over-all selectivity of the two stages is the product of the individual selectivities. If the response at a frequency far away from resonance is specified in decibels below that at resonance, the response of the two stages is the sum of the individual responses in decibels.

(2) The most important spurious response is the *image* frequency which combines with the local oscillator to produce a spurious i-f side band. If the local oscillator is higher in frequency than the desired channel by 1 mc, for example, the image response frequency is 1 mc higher, or 2 mc above the desired channel. If a strong carrier appears at the image frequency, it interferes with the desired station. The higher the intermediate frequency, the farther away the image is from the operating frequency and the greater its attenuation in the tuned circuits ahead of the mixer.

(3) Spurious responses of less importance than the image frequency can occur at many different frequencies where harmonics of the local oscillator beat with undesired signals (and their harmonics) to produce the i-f signals. Many other possible spurious responses such as these are troublesome when the receiver is operated in the vicinity of a strong transmitter. It is difficult in some instances to determine the cause of a particular response in a receiver.

b. Interaction of Oscillator and Signal Frequencies. In all mixers, a certain amount of coupling exists between circuits which introduce the r-f signal to the tube and those which introduce the local-oscillator signal. When the i-f is low compared with the operating frequency, the frequency of the oscillator and that of the mixer input circuit are very close together. If a strong signal appears in the mixer input circuit while the receiver is tuned to a weak signal on an adjacent channel, the stronger one tends to cause the oscillator to shift frequency and *lock in* with it. This results in failure of reception of the weak signal, and is called *pulling*. The degree of oscillator-frequency pulling is dependent on the i-f, on the coupling between the oscillator and input circuits, and on the basic stability of the oscillator itself. The condition is aggravated with avc applied to the mixer. In general, coupling between the oscillator and input circuits is greater at high frequencies, where oscillators tend to be less stable. Mixers have varying degrees of isolation between the oscillator and the r-f circuits. Therefore, the oscillator-mixer combination designed for use at high frequencies is strongly influenced by the degree of isolation. Converters, because of the association of oscillator and mixer in one tube envelope, generally are the worst offenders in regard to pulling.

c. Noise and Input Loading in Converters and Mixers.

(1) Like r-f amplifiers, the control grids of both mixer and converter tubes present a conductance which appears across the tuned input circuit. The amount of loading depends on the type of converter and the operating frequency. Where the r-f signal is introduced on an inner grid, with the local-oscillator signal on an outer grid, the loading is negative. This means that the resistive component of the conductance has a negative value at the operating frequency. The negative resistance tends to cancel out

some of the positive resistance of the tuned circuit. Therefore, its Q is raised, resulting in an improvement in image rejection. When the Q of a parallel resonant circuit is raised, its effective load impedance increases, and the output voltage from the r-f amplifier goes up. This improves the operation at high frequencies. In those mixers where the local-oscillator signal is injected on an inner grid, or on the same grid as the r-f signal, the input loading is positive. Here the action on the gain and image rejection is the reverse: Image rejection is lowered because of the lower Q, and the r-f gain is diminished because of decreased plate load impedance.

(2) Like r-f amplifier tubes, converters produce shot effect and partition noise. Because many converters have a larger number of grids than r-f amplifiers, the noise figure is higher. Even when triode tubes are used as mixers the noise figure is much lower than for the same tube operating as an r-f amplifier. This additional noise is contributed by the local oscillator. Any irregularity in the frequency of oscillation produces an effective noise voltage, in addition to the thermal, partition, and shot-effect noises generated in the oscillator tube and circuit. Generally, in f-m equipment for the v-h-f band, sufficient r-f gain is used ahead of the converter to prevent this noise from being a serious problem. Above the v-h-f range, however, r-f amplifiers are no longer practical, and crystal mixers which have a certain amount of loss but contribute nothing to the noise level are used. Of course, the local oscillator still will add noise, so that it becomes the main source of additional noise.

(3) The noise produced in multigrid mixers and converters can be reduced to a low value, equivalent in some cases to that of a pentode amplifier, by the proper application of feedback at the operating frequency. This feedback usually is generated across a tuned parallel-resonant circuit in the cathode lead. It is adjusted to resonance below the operating frequency so that its reactance is primarily capacitive. Variations in cathode current then appear as voltage variations across this tuned circuit. These voltage variations aid the voltage at the grid, effectively increasing the signal without adding to the noise. Therefore, the over-all noise figure is reduced. To prevent instability and oscillation, some of the signal frequency output from the plate circuit is returned to the grid through a neutralizing circuit. This does not interfere with the feedback that reduces noise. Similar arrangements can be used with crystal mixers to cancel out the noise introduced by the oscillator.

d. Oscillator Radiation.

(1) The local oscillator in an f-m receiver generally operates at a fairly low level, producing only microwatts of power. However, even this low power, if allowed to reach the antenna, can cause serious interference to receivers operating in the vicinity. The position of a receiver with a radiating oscillator can be located by unfriendly forces through triangulation. Therefore, the suppression of oscillator radiation is important in military receivers. The mixer is most critical in this respect. Oscillator power can reach the antenna back along the path through which the received signal comes. The coupling of the oscillator and signal sections of the mixer influences this strongly. Where the oscillator voltage is injected on the same grid as the signal voltage, the condition is especially acute. In those mixers where the oscillator is applied to a separate grid, the coupling depends on the tube capacitance. Part of the oscillator voltage reaches the signal-tuned circuit through direct or stray-capacitive coupling. Another portion reaches

it through the electron stream of the tube.

(2) The suppression of oscillator radiation for a particular tube takes place primarily in the tuned circuits of the mixer input and the r-f amplifier. At low frequencies, where outside noise levels are far in excess of internal tube noise, the r-f amplifier is used to attenuate the local-oscillator signal applied to the antenna. Since the selectivity is greatest far from resonance in a tuned circuit, higher intermediate frequencies result in better local-oscillator suppression. However, where crystal mixers are used without r-f stages, as in u-h-f (ultrahigh-frequency) operation, the radiation must be eliminated entirely in the input circuit. The very-high intermediate frequency required to allow for lowered selectivity makes the design and operation of i-f amplifiers difficult. In general, doubling the intermediate frequency reduces oscillator radiation by a factor of 4. In vacuum-tube mixers, if the i-f is close to the operating frequency, the load presented to the oscillator voltage through the capacitance of the tube is high, and the radiated voltage increases considerably.

e. Tracking of Mixer and Oscillator Circuits.

(1) When a superheterodyne receiver is tuned over a band of frequencies, a constant frequency difference must be maintained between the oscillator frequency and the incoming signal. When this difference is maintained properly, the oscillator tuning is said to *track* with the mixer tuning. The frequency of a tuned circuit will vary with changes in either the capacitance or the inductance. From the formula for resonance, it is clear that the frequency decreases as the square of the increase in capacitance or inductance. Since the oscillator generally is tuned to a frequency higher than that of the mixer, if variable capacitors are used in the oscillator and mixer circuits, the range of capacitance change of the oscillator must be restricted. The same change in capacitance in the oscillator as in the mixer would result in a larger oscillator frequency change at the higher frequencies, where a small change in capacitance is more significant. Restriction of the upper range of the oscillator tuning circuit is accomplished by a small capacitor in parallel with the main tuning capacitor. It effectively sets the minimum capacitance and consequently the maximum frequency of the circuit. Similarly, another much larger capacitor in series with the tuning circuit reduces the total effective capacitance so that the oscillator frequency maintains the required dif-

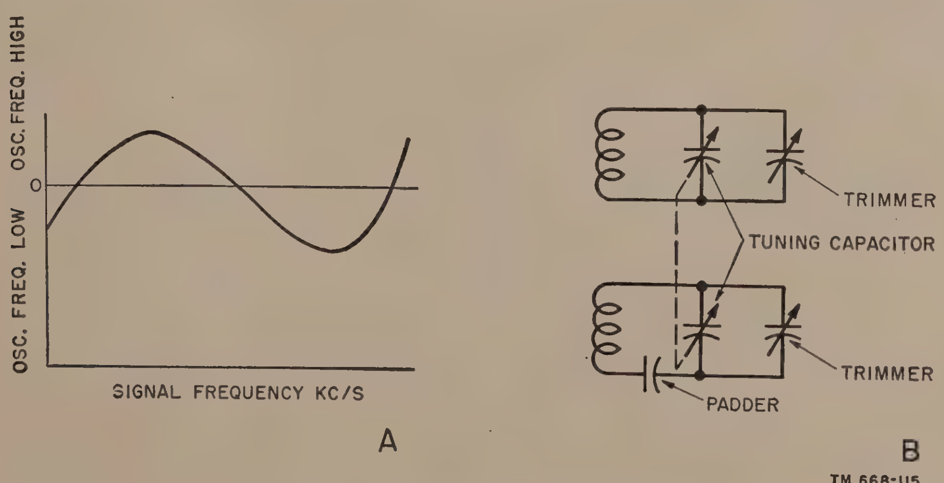


Figure 117. Tracking circuit and error curve.

ference from the mixer at the lowest frequency. By suitable adjustment of the inductance of the oscillator coil, the capacitance of the parallel capacitor (the *trimmer*), and the capacitance of the series capacitor (the *padder*), exact tracking can be obtained at three frequencies over the tuning range. Errors in tracking can be shown graphically, as in A of figure 117, where they are plotted against operating frequency. This type of error curve is typical for the circuit in B.

(2) When the inductance, rather than the capacitance, is varied (permeability tuned), the tracking is obtained either with variable inductances connected in series and parallel with the main inductance or with small trimmer and padder capacitors as before. Many double-conversion receivers require elaborate tracking arrangements. Special tuning capacitors may be used whose oscillator sections have differently shaped plates as compared with the mixer and r-f section. In this way perfect tracking can be obtained over the entire range without complex adjustment. This arrangement is not suitable, however, in a multirange receiver, since the frequency difference required between the oscillator and the mixer becomes proportionately smaller relative to the increased operating frequency. A coil can be wound with variable-pitch winding so that permeability tuning will give a constant frequency difference between oscillator and mixer. This method is used widely in the first i-f section of double-conversion receivers, with ganged-capacitor tuning in the second i-f section.

64. Mixer Tubes and Circuits

a. Diode Mixers.

(1) Diode mixers are used at the extreme upper end of the v-h-f range, where ordinary triode tubes are unsatisfactory. The diodes used have extremely small internal dimensions with close spacing between plate and cathode. A typical circuit using such a mixer is shown in figure 118. Diode mixers function not only with the oscillator fundamental, but also with harmonics. Therefore, oscillators operating at lower frequencies can be used. From the diagram, it is apparent that there is relatively little isolation between the oscillator and the r-f and antenna circuits, causing considerable oscillator radiation.

(2) The voltage from the oscillator is rectified, causing d-c to flow in the plate circuit. Optimum operation usually is obtained with d-c currents of approximately 200 to 500 microamperes. The parallel resonant circuit in the cathode is tuned to the oscillator fre-

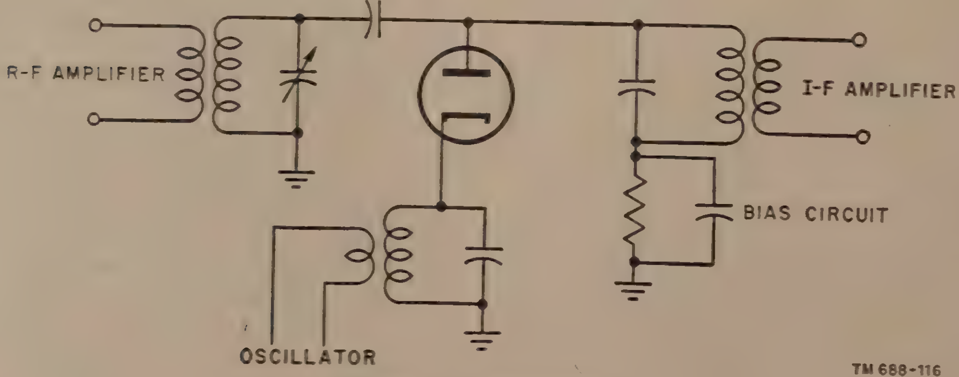


Figure 118. Diode mixer.

quency, with the capacitor made sufficiently large to act as a cathode bypass at the r-f signal frequency. The heater-cathode capacitance in most tubes is sufficient at high frequencies to render the oscillator injection circuit inoperative if not isolated. Therefore, r-f chokes always are required in the heater circuit with this type of mixer.

(3) Since the diode cannot have any conversion gain, the noise figure of this type of mixer depends entirely on the i-f amplifier which follows it. The diode produces a shot-effect noise, and noise, of course, is introduced by the oscillator injection. Therefore, the over-all noise figure of this type of mixer is not as good as in a crystal diode, where no filament and consequently no shot effect are involved.

b. Triode Mixers.

(1) Two triode mixer circuits are shown in figure 119. In A, the oscillator signal is injected along with the r-f signal at the grid; in B, cathode injection is used. As far as the performance of the mixer is concerned, there is little to choose between the two methods of oscillator injection, except that the grid loading with cathode injection is slightly greater. However, cathode injection gives better oscillator stability, since the load presented to the oscillator has a low impedance. The oscillator signal can be taken from a low-impedance point where varying load does not affect the operating frequency.

(2) The conversion transconductance is about 28 percent of the transconductance of the same tube operated as triode amplifier. If the transconductance at zero bias is reasonably high, the equivalent noise resistance of the tube is low, and the over-all noise is low. In general, the noise figure for a triode mixer is about the same as that for a single pentode amplifier. By the application of proper feedback, however, the noise from the oscilla-

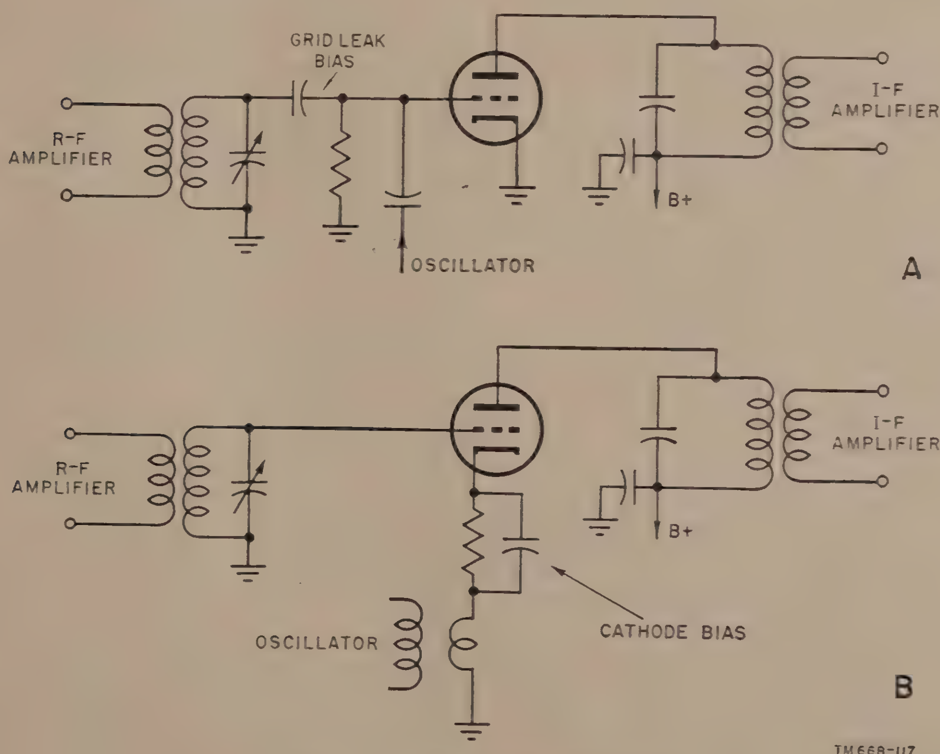


Figure 119. Single-ended triode mixers.

tor can be reduced so that the overall noise is no greater than that for the same tube used as an amplifier. It is possible to connect the tube as a grounded-grid mixer with cathode input and in this way obtain operation at higher frequencies than would otherwise be possible.

(3) If the operating frequency and the intermediate frequency are close together, the plate-load impedance becomes sufficiently great to permit tuned-grid tuned-plate oscillation. This causes instability, and the stage must be neutralized by conventional means. Furthermore, the capacitive plate load reflects a considerable resistive load across the mixer input, reducing the gain and the image rejection.

c. *Dual-Triode Mixers.*

(1) The push-push grounded-cathode circuit is used widely at v-h-f and functions effectively up to 600 mc. The oscillator signal usually is injected into the grid by means of a small inductive coupling loop (fig. 120). The local oscillator and input signal are completely cancelled in the plate circuit, which improves the signal-to-noise ratio and adds to the stability. The actual operation is identical with that of the balanced modulator, which it strongly resembles.

(2) Another dual-triode mixer that is useful at high frequencies is the cathode-coupled circuit shown in figure 121. The oscillator signal is injected through a cathode-follower section. The oscillator is isolated from the input circuit and from the mixer tube, which results in improved stability. There is almost complete elimination of any tendency of the oscillator frequency to shift with variation in the strength of received signal (pulling), as most other mixers would do with poor oscillator signal-grid isolation.

d. *Pentode Mixer.*

(1) The pentode shown in figure 122 is one of the most frequently used mixers in f-m equipment for the v-h-f

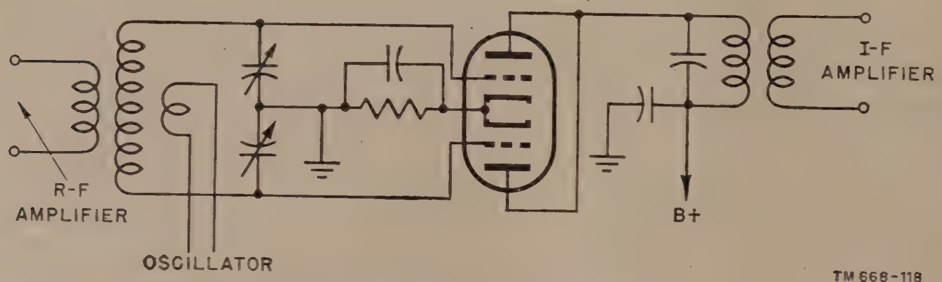


Figure 120. Push-push triode mixer.

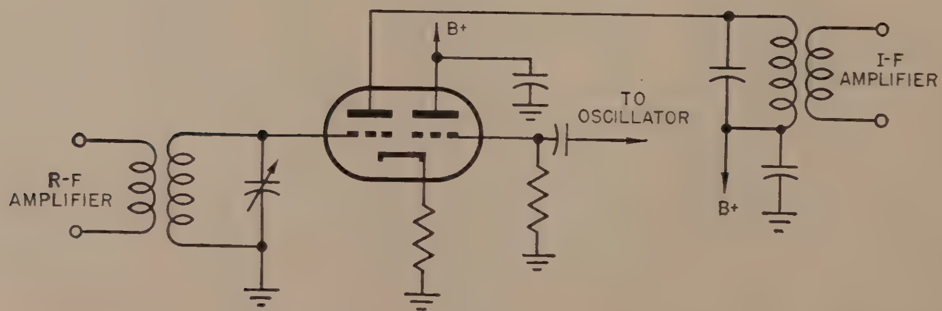


Figure 121. Cathode-coupled mixer.

band. At low frequencies, where the screen grid is effective, the mixer provides good isolation between the input and output circuits. This means reduced input loading and elimination of possible instability. The oscillator and signal voltages usually are applied to the signal grid. In this way, a noise figure is obtained which exceeds that of a normal pentode amplifier, but which is much lower than in any of the multigrid mixers.

cathode inductance, and cathode injection will lower the voltage gain of the input circuit and also the noise performance. The stability of the oscillator, however, is improved at very-high frequencies, where a low-impedance oscillator load is needed. Unless the oscillator and mixer are loosely coupled, interaction and pulling become severe. Interaction of the oscillator and the signal is greatest when they are both on the same grid.

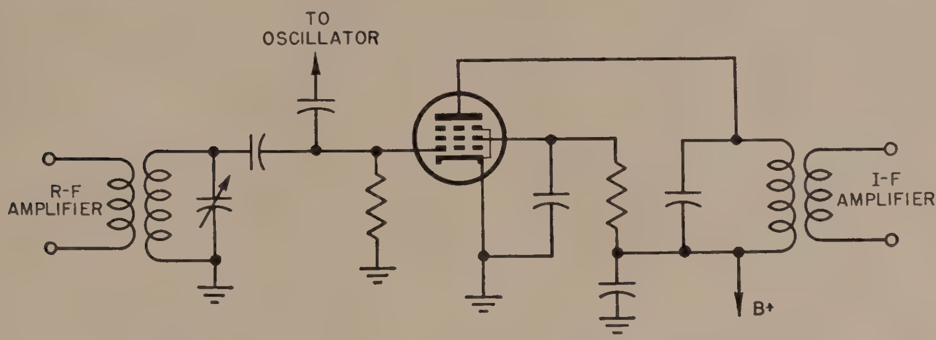


Figure 122. Pentode mixer.

(2) The pentode has an extremely high conversion transconductance and permits high voltage gain in the mixer stage. The equivalent noise voltage produced by the tube is twice that of a triode mixer of the same transconductance. Because of the high obtainable transconductance of pentodes, the over-all performance can exceed that of some triodes. Since the triode has a certain amount of coupling between grid and plate circuits, it is at a disadvantage compared with the pentode. At the signal frequency, the i-f circuit is capacitive, and this, because of Miller effect, results in a reflected low resistance in the grid circuit. The screen in a pentode effectively stops this loading.

(3) With a pentode, cathode injection of the oscillator signal is possible, but this mode of injection will increase the effective cathode inductance. Since the input load is proportional to the

Similarly, oscillator radiation becomes a greater problem; however, the high transconductance permits the use of small oscillator voltages, and radiation is not as great a problem as in a triode.

e. Multigrid Mixers.

(1) In superheterodyne receivers, many different tubes have been developed solely for use as mixers, but only a few of these are usable at v-h-f. In general, all multigrid mixers, which have more grids than a pentode, are divided into two classes—those in which the r-f signal is placed on the inner (control) grid, and those where the local-oscillator signal is on the inner grid. In the first category are the pentagrid mixer and the heptode. In the second are the pentagrid converters, which include the hexode and the octode. In addition to these tubes, there are those with a triode local-oscillator section in the envelope with

a pentode mixer. These do not differ from the separate mixers and oscillators covered in the preceding paragraphs.

(2) A cross section of a typical pentagrid mixer is shown in A of figure 123. The oscillator voltage is applied to the oscillator grid, and the r-f signal is applied to the signal grid, which is the third grid out from the cathode. The signal grid is shielded from the oscillator grid by a screen grid. The screen grid is connected to a second screen between the signal grid and the plate which serves to isolate the signal grid from the plate. A conventional suppressor follows the second-screen grid. A typical circuit using this tube is shown in B. It has fairly high gain, because the internal construction permits a high conversion transconductance.

(3) Because of the large number of grids, the noise performance is poor, but this is largely offset by the high gain. With suitable feedback in the cathode circuit, the noise figure can be brought down to that of a pentode. The major fault arises from coupling between the signal grid and the oscillator grid through the space charge of the tube. This reduces the gain and the conversion transconductance. A small amount of capacitive coupling exists even though a screen is placed between the two circuits, and this adds to the interaction between the oscillator and mixer sections. The effect of the capacitive coupling is opposite to that of space-charge coupling, and if a small capacitor is placed between the oscillator and the signal grids to add to the normal capacitance, the space-charge coupling can be neutralized at high frequencies. This improves the performance somewhat. The tube loads the tuned circuit in the signal grid negatively so that the performance of that circuit actually is improved. However, during the nega-

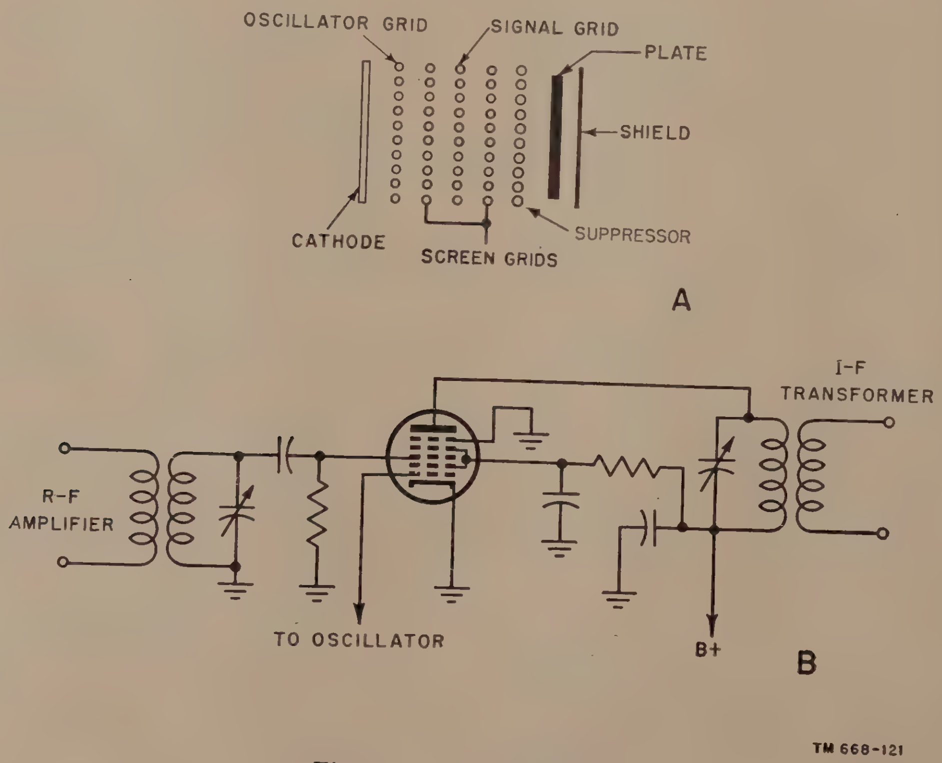


Figure 123. Pentagrid mixer.

tive portion of the oscillator cycle, the cathode can swing more negative than the signal grid. When this occurs, the signal grid draws current unless the oscillator grid is sufficiently negative to cut off the cathode current entirely. Therefore, suitable values of grid-leak resistance must be used to develop a high negative bias, or the oscillator voltage must be increased. The tube is used more often as a converter than as a mixer.

65. Converters

a. Converters are essentially mixer tubes where the oscillator grid in conjunction with the electron stream and the plate or screen acts as a self-contained oscillator. The oscillator, therefore, does not require a separate tube. The pentagrid converter is the principal type used, and a typical circuit is shown in A of figure 124. In a converter, the requirements of oscillator stability are most severe. Only those types can be used that have an internal construction which will allow the frequency of oscillation to be stable. The mixer part of the tube is conventional, but the oscillator consists of the inner grid, the cathode, and the screen grid. The screen acts as the plate, so that local oscillation takes place between elements forming a triode.

b. The converter is used widely in the second mixer-oscillator stage of double-conversion receivers, with a crystal oscillator as shown in B. It is a highly stable circuit with good gain at the low frequencies involved. The crystal oscillator is an ultraudion type, with the crystal acting as a tuned circuit between the screen and the control grid of the oscillator section, the screen acting as a plate. At the higher frequencies, the gain of the converter falls off unless sufficient capacitance is added between the oscillator and signal grids to neutralize the effect of space-charge coupling.

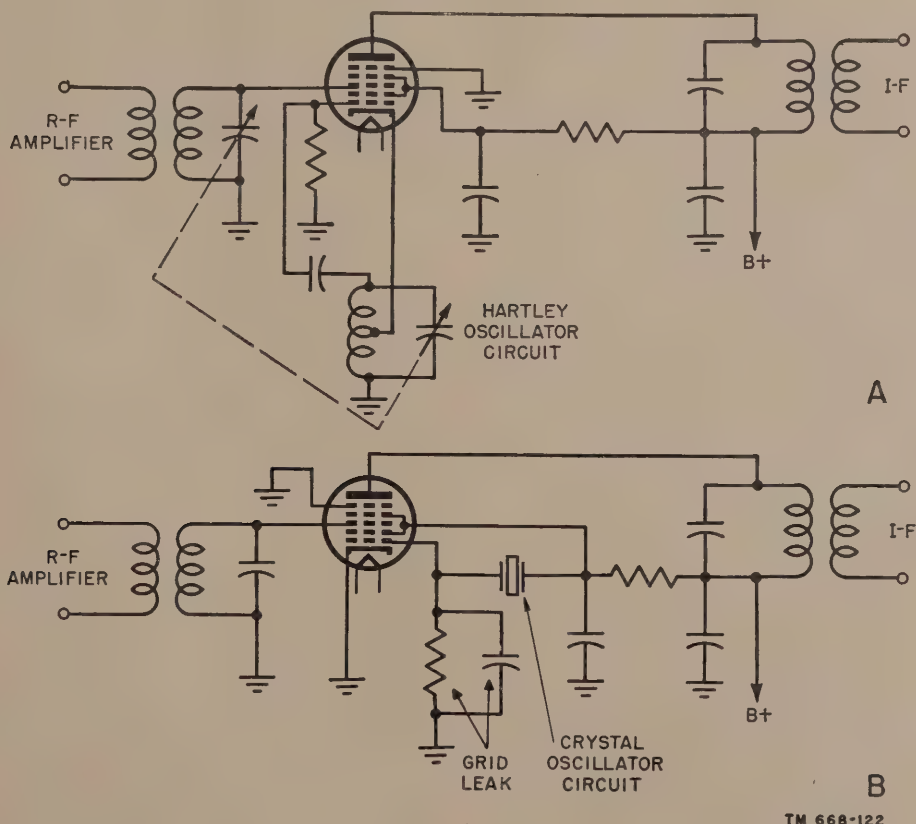


Figure 124. Pentagrid converters.

66. Crystal Mixers

a. General. At extremely high frequencies, vacuum tubes no longer function satisfactorily as mixers, and germanium or silicon diodes are used. The mixer circuit (fig. 125) is nearly the same as that of the diode mixer, although no bias or filament voltages are needed. The crystal operates because there is a much higher resistance to current passing in one direction than in the other. Therefore, the crystal is essentially a nonlinear device.

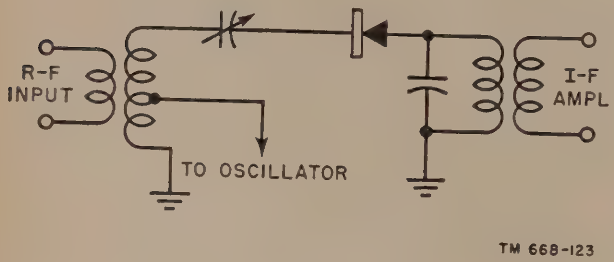


Figure 125. Crystal mixer.

b. Oscillator Injection. The performance of the crystal mixer depends on the uniformity and amplitude of the local-oscillator voltage injected. The oscillator voltage affects the matching of the r-f and i-f circuits and consequently the over-all noise performance. The average input impedance is approximately several hundred ohms, and the output impedance is about the same value, or slightly lower. However, they are not a constant and both input impedance and output impedance decrease as the rectified current produced in the crystal by the local oscillator increases. If ordinary methods of oscillator coupling with small capacitors are used, the injection of oscillator current varies widely, consequently changing the impedance. This affects adversely the noise performance and the conversion efficiency. Therefore, to provide uniform oscillator injection, equalizers are inserted in the circuit.

c. Conversion Efficiency. A crystal produces no voltage gain. However, the conversion loss is slight with well designed crystals. The effective loss in signal-to-noise ratio, when measured in terms of noise figure, also includes a factor resulting from the *excess temperature noise,* which is the additional oscillator noise measured by the rise in temperature needed to produce the same amount of noise with the same impedance level at the crystal input. The conversion efficiency is relatively high if sufficient oscillator injection is available. Power of approximately 500 microwatts usually is sufficient. This is much lower than that required by a vacuum-tube mixer.

d. Frequency Response and Noise Performance. The frequency response of a crystal is practically uniform from the low-audio frequencies up to the superhigh-frequency range. Germanium crystals are capable of withstanding higher voltages, whereas the silicon crystals have less noise and conversion loss. Since the crystals are used in ranges where r-f amplifiers are not practical, the conversion efficiency must be as high as possible with little noise. The over-all noise performance of an f-m receiver using a crystal mixer is determined almost entirely by the noise figure of the first i-f amplifier. In this respect, the i-f amplifier plays almost the same role as the r-f amplifier does at lower frequencies. It is necessary to use a high intermediate frequency with a crystal mixer to minimize oscillator radiation and excess temperature noise.

e. Crystal Operation at High Frequencies. The conversion transconductance of a crystal cannot be specified by a simple constant, since it depends on frequency and oscillator excitation. Because the spacing is extremely close, the capacitance between the fine wire in contact with the crystal surface in the holder is considerable despite the small area. No reliable prediction can be made of performance at high frequencies from measurements made at low frequencies. At these high frequencies, ordinary coils and capacitors are not satisfactory as circuit elements, so that it is common practice to use sections of transmission line, or waveguide. However, the operation of the crystal as a mixer is the same as any other nonlinear device, and at high frequencies, the over-all performance obtainable with good crystals approaches that obtainable at lower frequencies with vacuum tubes.

67. L-C Oscillators

a. General. The local oscillator of a superheterodyne f-m receiver generally operates at a high frequency. In a double-conversion heterodyne, however, the second local oscillator operates at a much lower frequency. The stability of the entire receiver is determined by the h-f oscillator since any change in oscillator frequency changes the effective tuning of the receiver. The principal requirement for high-frequency oscillators in f-m receivers is a high degree of stability in respect to thermal, mechanical, and electrical variations.

b. Lumped-Constant Circuits.

(1) Most oscillators used at very-high frequencies tend to be unstable unless great care is taken in the construction of the parts which determine the frequency. This instability stems from the large changes in reactance that take place with small changes in capacitance or inductance. The most satisfactory circuits are those in which the effect of the vacuum tube on the frequency of oscillation is least. This usually is accomplished through the use of the conventional Colpitts oscillator (fig. 126) or some modification of it. The conventional Colpitts oscillator in A obtains feedback through two capacitors in series, which are also effectively in parallel with the tube interelectrode capacitances. Since the interelectrode capacitances are small, large values of feedback capacitance make the effects of tube capacitance negligible.

(2) A maximum value of series capacitance, however, is determined by the necessary load impedance to be developed by the tank circuit in order to sustain oscillation. To avoid some of the disadvantages of using relatively low values of capacitance, the circuit in B often is used. The series capacitors are a part of the tank circuit, with a smaller series capacitor used to tune the oscillator. This arrangement permits stability over a wide range of variation in either tube or load conditions. Both circuits, however, have the disadvantage of a tuning range restricted by the size of the tuning capacitors. Therefore, even though stability is lower, conventional Hartley or Armstrong circuits are used where wide tuning range is necessary.

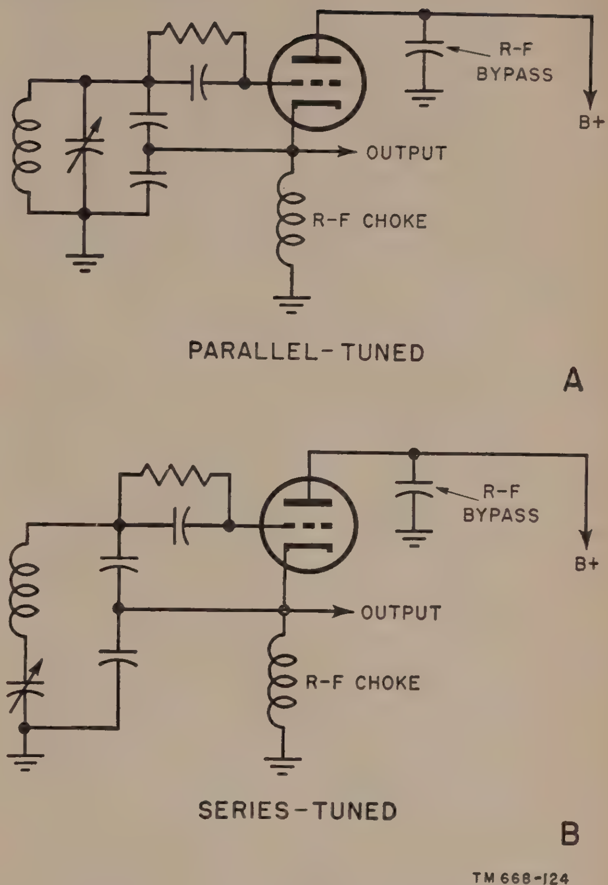


Figure 126. Two versions of Colpitts oscillator.

c. Causes of Drift.

(1) The causes of drift in lumped-constant oscillators operating at high frequencies can be classified electrically, mechanically, and thermally. Electrical drift takes place when the plate voltage, filament voltage, or any other tube potential changes. For most oscillators, the supply voltage enters into the determination of the frequency of

oscillation, although simplified theories of oscillator operation usually do not take this fact into account. This is because a vacuum-tube oscillator is a device whose characteristics vary with the supply voltage. The variations can be corrected by suitable voltage regulators for the plate and filament supplies. Mechanical drift arises when inductors and capacitors change their physical dimensions with vibration. This usually is overcome by using very rugged components and shock-mounting the unit.

(2) Thermal drift takes place because the values of capacitance or inductance change as the dimension of the components change with temperature. Low temperature-coefficient ceramic materials, stretched wire coils, temperature-compensating capacitors, and similar devices reduce the drift. Associated with thermal changes are humidity variations. Moisture-laden air has a different dielectric constant from that of dry air, and the capacitance of air capacitors changes with variations in humidity. To overcome this, a heating device can be located near the oscillator tank to keep the air dry, or the tank circuit can be hermetically sealed. Permeability tuning, which is much less susceptible to humidity, also can be used where other methods are not satisfactory. In all instances, the instability can be reduced by lowering the frequency of the oscillator and then using frequency multipliers to raise it to the desired value.

d. Distributed-Constant Oscillators. At frequencies that are too high for ordinary lumped-constant oscillators, the tuned circuits are made of sections of transmission line. A shorted coaxial line serves as an inductance if its electrical length is less than a quarter-wavelength at the operating frequency. These transmission-line sections have very high Q and, if properly designed, are stable mechanically. Receiver oscillators using transmission lines tuned with variable capacitors are common in v-h-f equipment. The inductance is distributed along the entire length of the line, and therefore is called a distributed constant. The operation of these oscillators is the same as for those used at low frequencies except that the tuned circuit does not have a lumped inductance.

e. Second-Conversion Oscillators. In the second conversion circuit of double-conversion superheterodynes, the local oscillator can be of any conventional type. In some double-conversion receivers, the tuning is done in the second oscillator because it is easier to obtain stability at low frequencies.

68. Crystal-Controlled Oscillators

a. General. Crystal oscillators are used in f-m receivers because of their extremely high stability. Crystal oscillators used in second-conversion circuits operate at low frequencies with output at the frequency of the crystal. When crystals are used in the first-conversion circuit, the crystal oscillates at its fundamental frequency, but harmonic output is obtained through a harmonic generator. There are also circuits that cause the crystal to oscillate directly on a mechanical harmonic. For low-frequency circuits, the choice of circuit depends on its simplicity and economy of parts. In the high-frequency converter, the requirements are more complex since there is conflict between requirements for high harmonic output, low subharmonic output, stability, and low-crystal current.

b. Low-Frequency Crystal Oscillator.

(1) The crystal is the equivalent of a high Q resonant circuit and controls the frequency of an oscillator precisely. Each crystal oscillator circuit has its counterpart in an L-C oscillator. For example, the triode-crystal oscillator in A of figure 127, is the equivalent of the tuned-plate, tuned-grid oscillator. In B, the crystal oscillator circuit is the equivalent of the ultraudion oscillator. In both circuits, the output voltage is at the frequency of the crystal. The tuned-plate, tuned-grid arrangement is used rarely because precise tuning of the output circuit is necessary to maintain oscillations. The ultraudion circuit oscillates because of the feedback provided by the voltage

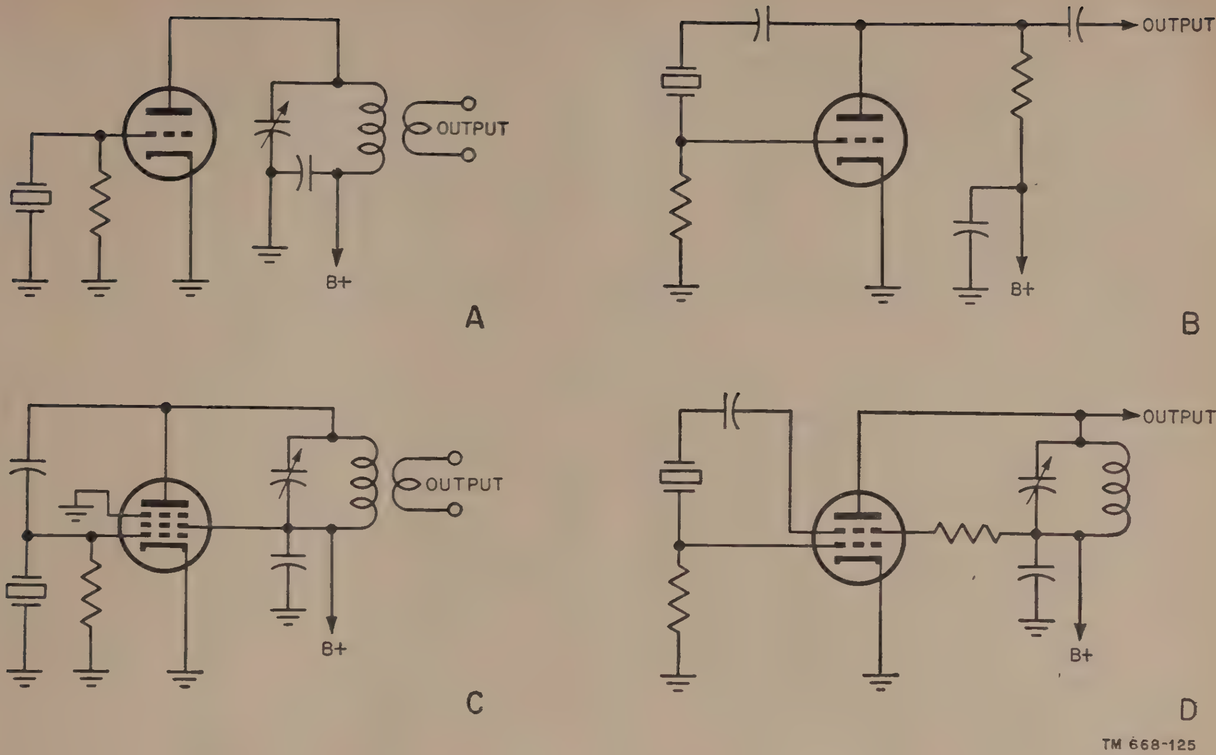


Figure 127. Low-frequency crystal oscillators.

divider formed by the capacitance from the grid to the plate and the cathode. It is perhaps the simplest of all crystal oscillators, since it requires only resistors and capacitors.

(2) Other crystal oscillators whose output voltage is of the same frequency as that of the crystal are shown in C and D. These are designed to operate with tetrodes and pentodes. The circuit in C is the same as the tuned-grid, tuned-plate in A, except that the triode is replaced with the pentode. The circuit in D is similar to the ultraudion in B, but it uses a tetrode.

c. Harmonic-Oscillator Circuits.

(1) Where the local oscillator of the f-m superheterodyne operates at a very-high frequency, it is not possible to use a crystal that oscillates directly at this frequency. One solution for obtaining output in the v-h-f range is to use an oscillator which generates a strong output on harmonics of the crystal fundamental, and then to use separate frequency multiplication to obtain the desired frequency.

(2) Figure 128 illustrates three harmonic-oscillator circuits for pentodes. The cathode, control grid, and screen grid form a triode oscillating at the crystal frequency. These oscillations are rich in harmonics and the plate circuit is tuned to the desired harmonic. A high impedance is presented to the selected harmonic and a considerable output voltage is developed across this load. The three circuits differ in the basic type of oscillator; that is, in control grid-screen, grid-cathode connections. In A, the crystal is inserted between the control grid and the screen grid. In B, it is inserted between the cathode and control grid.

(3) In C, a resonant circuit in series with the cathode is tuned midway between the crystal frequency and the second harmonic. Therefore, it is inductive at the crystal frequency. In conjunction with the capacitance between grid and

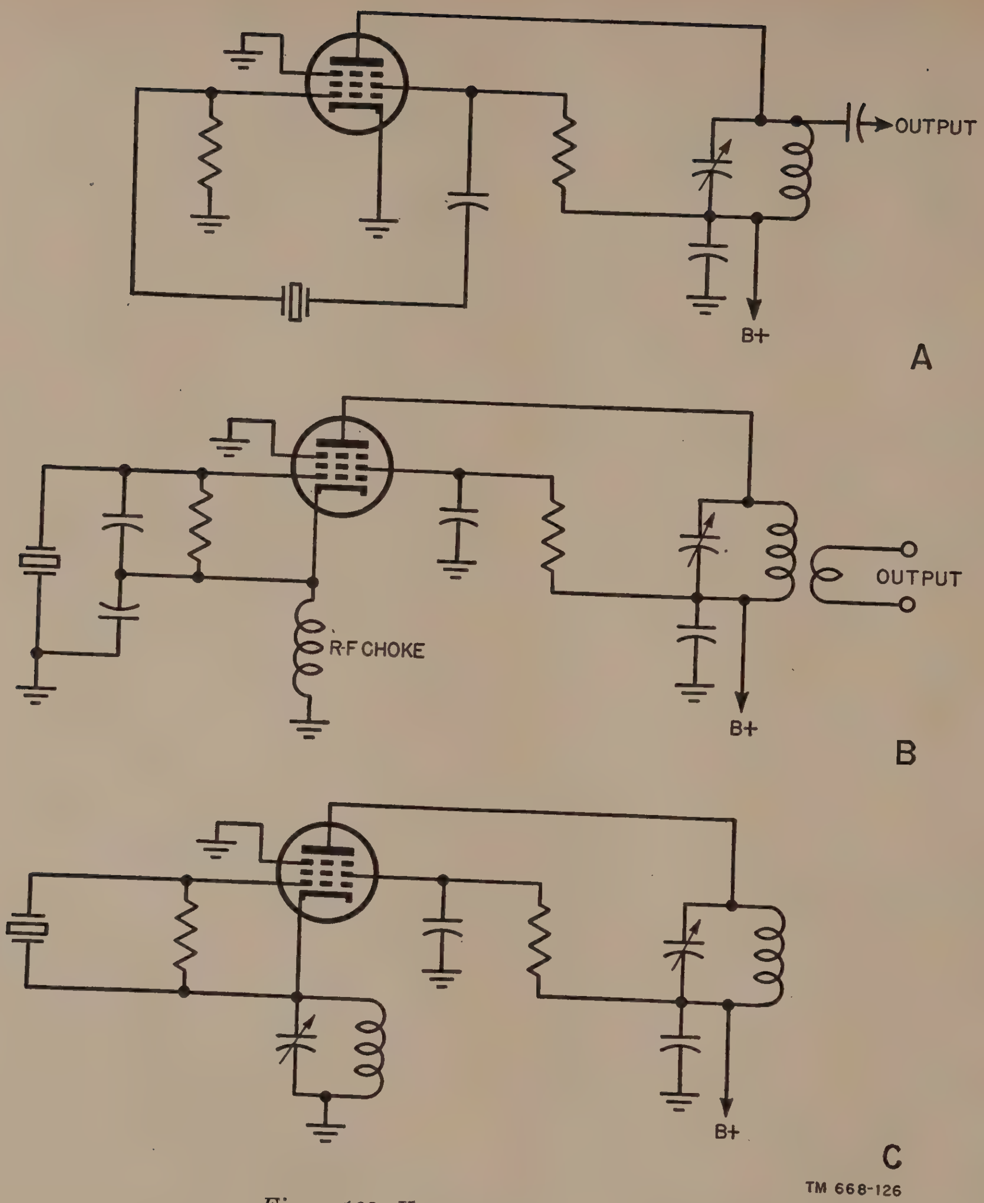


Figure 128. Harmonic crystal oscillators.

cathode, it acts as a voltage divider providing the required feedback for oscillation. Of these three circuits, that of C has the greatest harmonic output and the highest stability when it is used with a well-screened pentode. The second circuit is practically independent of variations in tube characteristics, if the feedback capacitors are large enough.

d. Overtone Oscillators.

(1) The output of all of the harmonic oscillators contains components at frequencies other than the fundamental. Unless they are separated from the mixer by a sufficient number of tuned circuits, serious difficulty with spurious responses can arise. These disadvantages can be overcome with special circuits and crystals. Instead of oscillating at their fundamental frequency, crystals can be made to oscillate on other frequencies very close to odd harmonics of their fundamental.

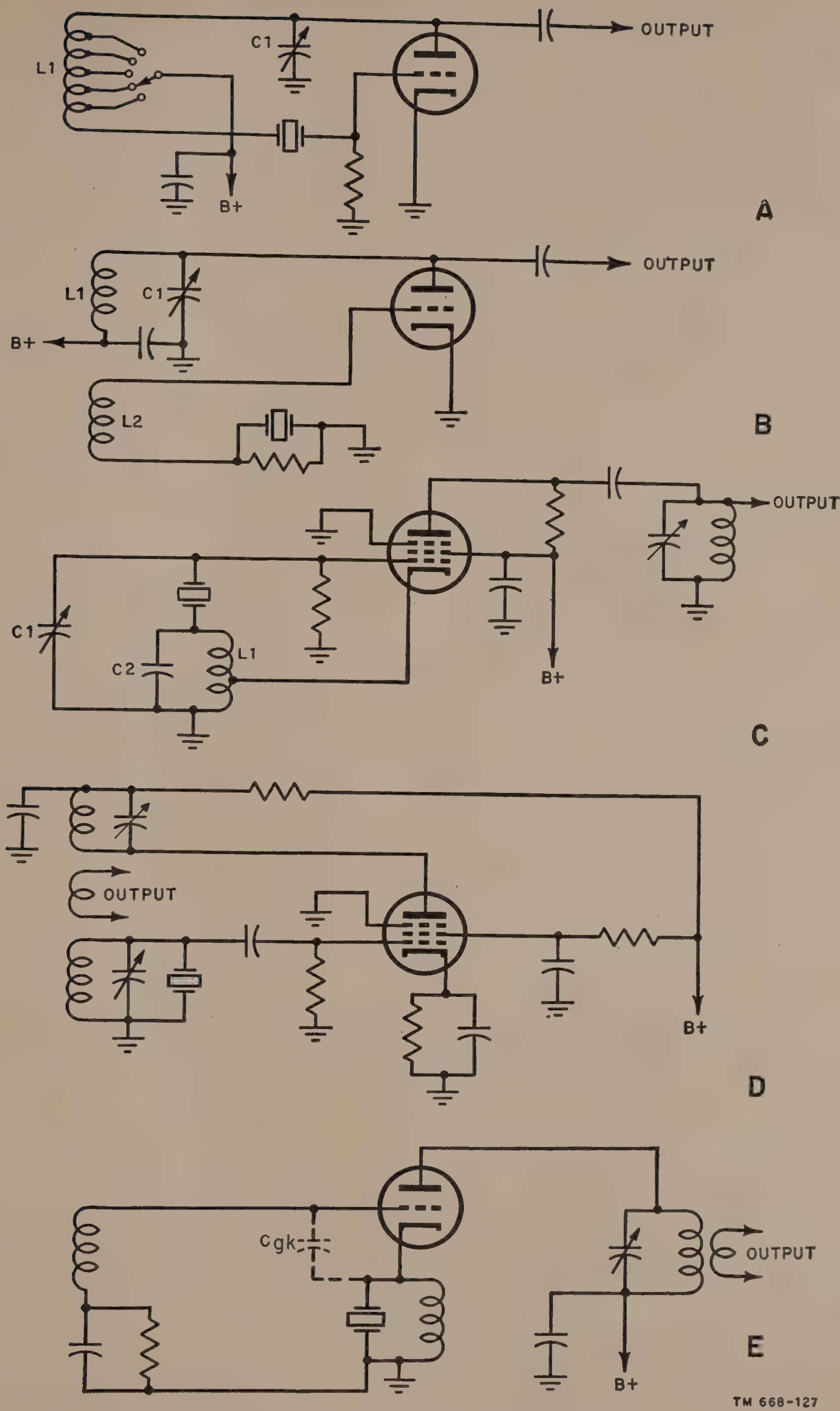


Figure 129. Overtone oscillators.

These frequencies are its mechanical harmonics. Oscillators using this principle are called *overtone oscillators*. Special types of circuits must be used for this purpose. Their advantage lies in negligible output at frequencies lower than the desired frequency, which reduces the number of possible spurious responses considerably.

(2) All of the overtone oscillator circuits (fig. 129) incorporate some form of frequency-selective feedback unlike that in a normal crystal oscillator. An additional resonant circuit is used to feed back energy at the frequency of the desired overtone, so that oscillation takes place only at the desired frequency. The frequency-selective feedback must be adjusted carefully, so that oscillation is caused only by the crystal and not by the feedback circuit.

(3) In A, a simple ultraudion triode oscillator is modified so that the frequency of feedback is controlled by the tuned circuit formed by $L1$ and $C1$. The amount of feedback is varied by changing the tap on $L1$. A second version of the same oscillator is shown in B, where the amplitude of the feedback is controlled by the coupling between $L1$ and $L2$. These circuits are especially suitable for operation of the crystal on the third overtone, where the frequency of feedback is not very critical. Power output at the fifth overtone is poor, and the feedback adjustment becomes fairly critical when ordinary crystals are used.

(4) The overtone oscillator in C is capable of extremely stable operation, although it works only with crystals designed for overtone service. Feedback is controlled by the position of the cathode tap on $L1$ and by the setting of capacitor $C1$. The tuned grid circuit is set for resonance at the frequency of the desired overtone. Therefore, the capacitance of the crystal and its holder is also part of the resonant circuit, which is similar to that of the Hartley oscillator. This circuit has good output, especially at higher overtones, and the plate circuit can be tuned to a harmonic of the overtone, producing further frequency multiplication.

(5) An overtone oscillator which uses a high-gain pentode is shown in D. Feedback is obtained by magnetic coupling. The crystal actually is resonated slightly above the desired overtone. This makes the entire grid circuit equivalent to a high-impedance parallel-resonant tank, which easily picks up a regenerative signal from the plate circuit and produces oscillation. At very high overtones, the capacitance of the grid circuit is sufficient to resonate with the inductance in the plate circuit. This circuit is capable of producing moderate amounts of output on extremely high overtones. Operation on the twenty-ninth overtone has been obtained. This represents an important saving of frequency multiplier stages, as well as freedom from spurious responses attributable to the subharmonics of doublers and triplers.

(6) The circuit in E is somewhat different from those above. The input and output circuits of the triode are tuned to the same frequency, although the grid is much more broadly resonant than the plate. This is because of the high ratio of inductance to grid-cathode capacitance. The crystal-holder capacitance is resonated in the cathode circuit with an inductor at the desired overtone frequency. The only frequency at which oscillation can take place is that at which the crystal goes through series-resonance, producing a low impedance from cathode to ground. To prevent self oscillation between grid and plate circuits, the inductance in the cathode circuit must be adjusted carefully.

69. Front-End Design

a. Since few f-m receivers are designed to operate on a single frequency, means must be provided for tuning them over the required range. The mechanisms used, as well as the circuits are designed to operate in the r-f amplifier, mixer, and oscillator stages and are known as the *front end* of the receiver. In many instances, operation is required over a considerable portion of the v-h-f spectrum, and the requirements placed on the tuning circuits become severe. Some form of automatic tuning also is desirable.

b. Since all tuned circuits use inductance and capacitance in one form or another, all front ends are designed to vary one of these basic circuit elements. At frequencies low enough for lumped constants, variable capacitors or variable inductors are used. Different means are available for adjusting these elements, such as powdered-iron slugs, tapped coils, or variable capacitors. At higher frequencies, distributed-constant devices, such as transmission lines, are used. The length of the transmission line is made either electrically or mechanically adjustable. Where crystal control is used, switching circuits have been evolved to change from channel to channel.

70. Automatic Tuning Mechanisms

a. Detent Selector. One of the simplest of all automatic tuning selectors is the detent switch. This device enables a shaft used for tuning to be set repeatedly at the same position (fig. 130). The actual tuning usually is accomplished by a variable capacitor connected to the shaft. A gear and dial-drive arrangement permits selection of frequency groups comprising different channels, and also selection of individual channels within each group.

b. Mechanical Push-Button Systems. A device that rotates the shaft of a variable capacitor to preset positions is shown in figure 131. The buttons are attached to notched rods resting on gears. When the button is depressed fully, the rod pushes the gear to a point where it can rotate no farther. The gear, in turn, rotates a variable capacitor to the desired position.

c. Electrical Push-Button Systems. Electrical push-button systems are manually operated electrical switches which connect various trimmer capacitors in turn across the tuned circuit. The setting of these trimmers determines the frequency to which the receiver is tuned. Included in this arrangement is a mechanical

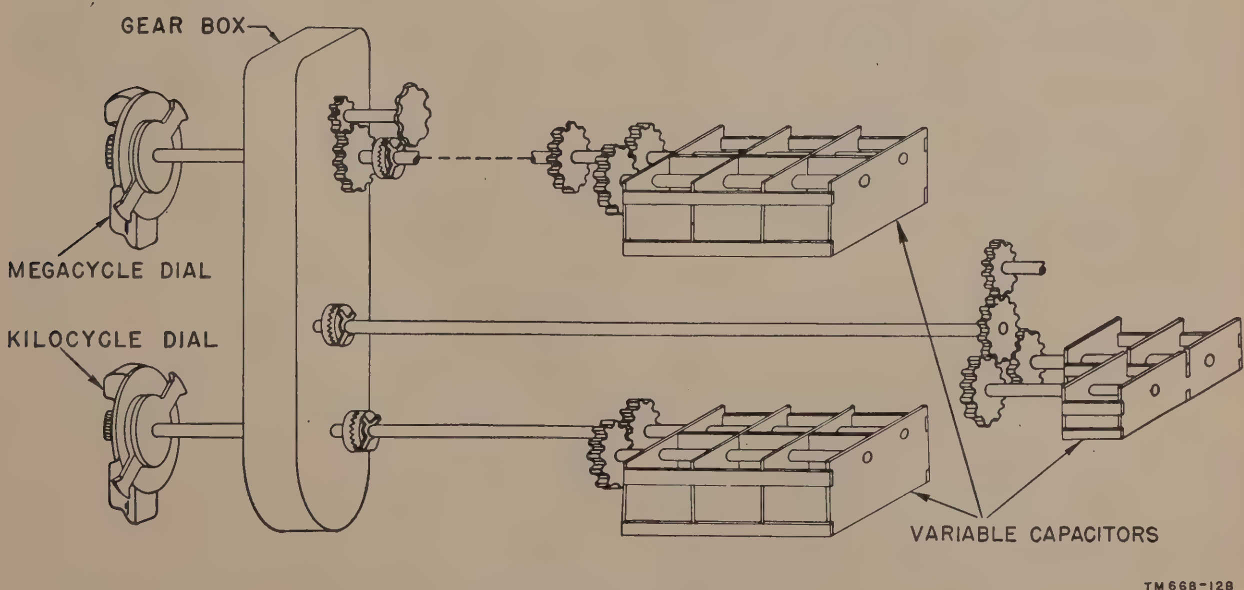


Figure 130. Detent mechanism.

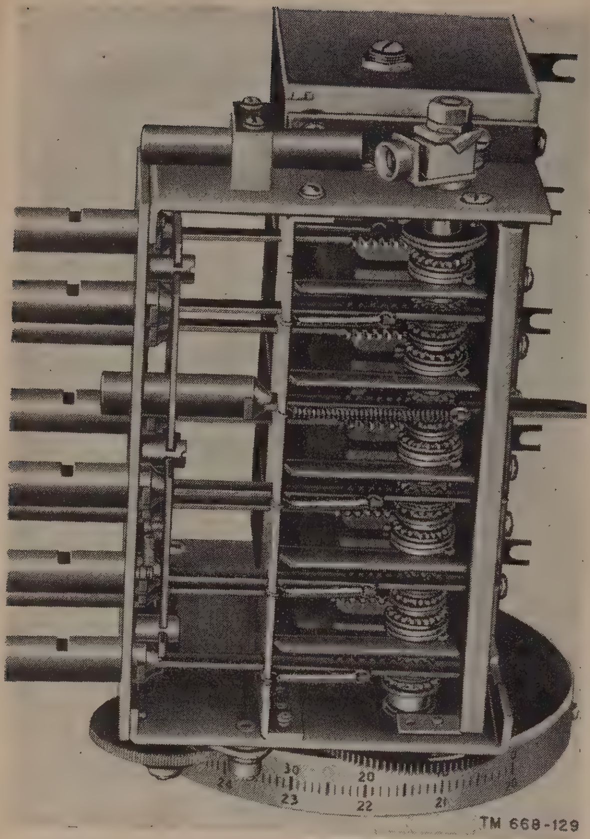


Figure 131. Mechanical push-button system.

latching gear that makes it impossible to depress two buttons simultaneously. In addition, the latching bar also disengages any button when another is pressed. The last button pressed stays engaged on the latch bar, indicating the channel in use. Because of excessive r-f losses in the complicated structure of the latching switch, it is not as satisfactory at v-h-f as are the mechanical positioning systems. However, up to about 100 mc, the performance is satisfactory if there is sufficient shielding.

d. Switch and Turret Tuners.

(1) A rotary selector switch used with preset coils to form a front-end circuit has many advantages over the ones previously mentioned. A rotary switch with several decks can connect in turn the appropriate reactors for up to approximately 12 channels. Each preset channel can be adjusted for optimum performance without much interaction from the other reactors if sufficient shielding is used. This system has good noise performance as well as low circuit capacitance, which is especially valuable at high frequencies. However, the limitation in the number of channels, as well as the necessity for very elaborate equipment, limits the use of this arrangement. A more flexible variant combines the switch tuner with a variable capacitor. The switch selects the appropriate coils and trimmer capacitors, and the variable capacitor provides fine tuning over the range of the particular coil.

(2) One of the most efficient of all the multichannel selectors is the so-called *turret* tuner. In this device, all of the channels have separate sets of inductors and capacitors which are mounted in a drum, or turret. The efficiency is high because the leads between the tuning elements and the tubes can be made short, permitting high circuit inductance and low over-all capacitance, which in turn means high tuned-circuit gain. As with all preset channel tuners using fixed electrical elements, it is difficult with the turret tuner to change a whole group of channels. This disadvantage can be overcome by providing an auxiliary variable capacitor or inductor which permits tuning through a considerable range.

71. Manual Tuners

a. Fixed-Frequency Equipment. In receivers where there is no need to change the band of frequencies, and where the total tuning range is small, separate coils and capacitors which are tuned manually are provided for each stage. An ordinary socket serves for the crystal oscillator, and different crystals are plugged in. The tuning of the r-f amplifier and mixer stages is accomplished with screw-driver-adjusted trimmer capacitors. At higher frequencies, where coils and capacitors no longer provide the proper amount of reactance, small setscrews which vary the length of a tuned transmission line are used in a similar fashion. Other arrangements, less frequently used, involve devices that are

combinations of inductors and capacitors and are varied with a setscrew.

b. Variable High-Frequency Equipment. The receiver shown in figure 132 has coaxial lines made of tubing. These are inclosed in square cans and are tuned electrically by a ganged variable capacitor. Another arrangement has been tried, using parallel-tuned lines that are actually parts of the capacitors. So-called *guillotine* tuners, which are essentially variable inductances, also are used. The parallel line is shorted by a bar, giving the appearance of a guillotine. Another tuned-line arrangement uses flat-plate transmission lines, with variable capacitors formed by small metal disks for adjustment of resonance range.

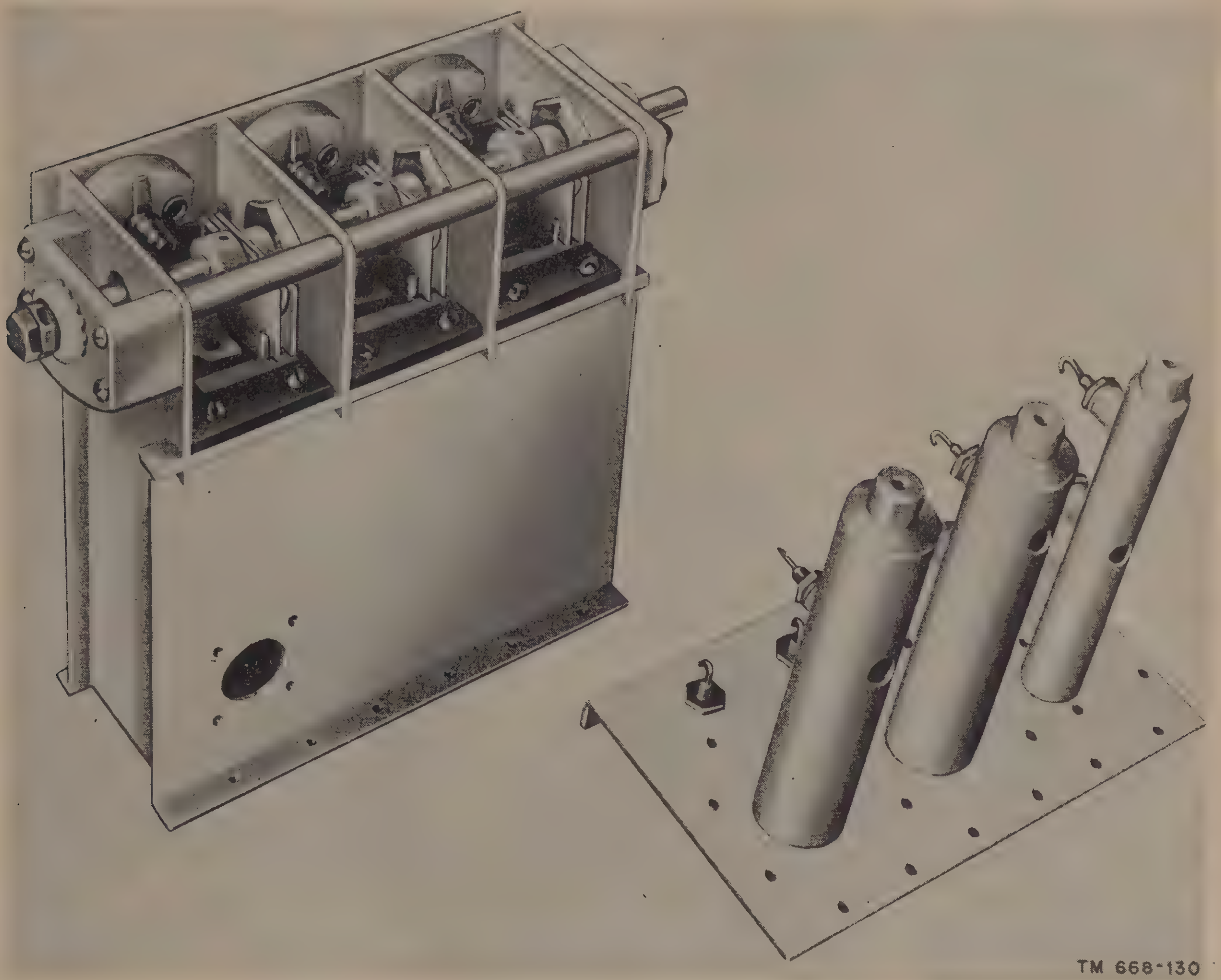


Figure 132. Manual tuning of coaxial lines.

Section VI. I-F AMPLIFIERS AND LIMITERS

72. Function

a. General. The front end of the f-m receiver produces an output signal usually in the microvolt range, whereas the detector requires signals in the order of several volts for its operation. Therefore, the i-f amplifier must perform most of the voltage amplification. The i-f stages must amplify the signal between 100,000 and 1,000,000 times. They also must introduce sufficient selectivity to discriminate against stations operating in the adjacent channel, and yet

have response sufficiently broad that the outer side bands of the f-m signal are not distorted.

b. Selectivity. The selectivity of an i-f amplifier is a measure of the total response of all of its tuned circuits. To obtain the required gain with adequate bandwidth, two or three i-f stages usually are needed. The circuit of a three-stage i-f amplifier is shown in figure 133. Each of the eight tuned circuits from the output of the mixer to the input of the detector has a response curve of the familiar bell shape. The response of the entire amplifier is the product of the responses at each point of the curve for each tuned circuit. The broadness of the peak of the curve for each tuned circuit depends on the Q. If the Q is 100 at a frequency of 10 mc, the peak of the curve is 100 kc wide between points where the voltage falls off to .707 of the peak value, or 3 db down. In general, the response of a tuned circuit is given approximately by the relation,

$$\text{bandwidth in kc} = \frac{\text{center frequency in kc}}{Q}.$$

When two tuned circuits are used, the response of the circuit is sharpened. or example, the over-all bandwidth of the circuit is 100 kc, and the response, by definition, is 3 db down 50 kc to either side of the center frequency. Applying the output of this circuit to a similar one means that a signal 3 db down at the input is reduced another 3 db, or a total of 6 db. Neglecting the mutual inductance, the response at 50 kc from the center frequency for an amplifier with eight tuned circuits is

$$8 \times 3 \text{ db} = 24 \text{ db down.}$$

This illustrates how the bandwidth is narrowed by increasing the number of tuned circuits. As the bandwidth is narrowed, the selectivity is increased.

c. Selectivity Requirements.

(1) In low-frequency f-m equipment, the channel spacing between adjacent transmitters can be as low as 50 kc. To prevent interference between channels at the receiver, considerable selectivity is required; however, increasing the selectivity of a stage increases the amplification of that stage. Because of difficulties resulting from feedback and instability, there is a practical limit to the amplification that can be obtained from an i-f amplifier. It seldom is possible to obtain high selectivity in a single i-f section. For good image rejection a high i-f is needed, which further restricts the selectivity that can be obtained. The double-conversion receiver separates the i-f section into two sections operating on different frequencies and helps reduce instability. The high i-f permits the necessary image rejection, and the low i-f provides the necessary selectivity.

(2) For higher frequency operation, where wide-band f-m is used, double conversion is not needed, since the adjacent channel-rejection requirements are less severe. However, wide deviation means that the selectivity of the i-f amplifier must be sufficiently broad to handle the full swing and still provide high gain and adequate rejection of off-channel signals.

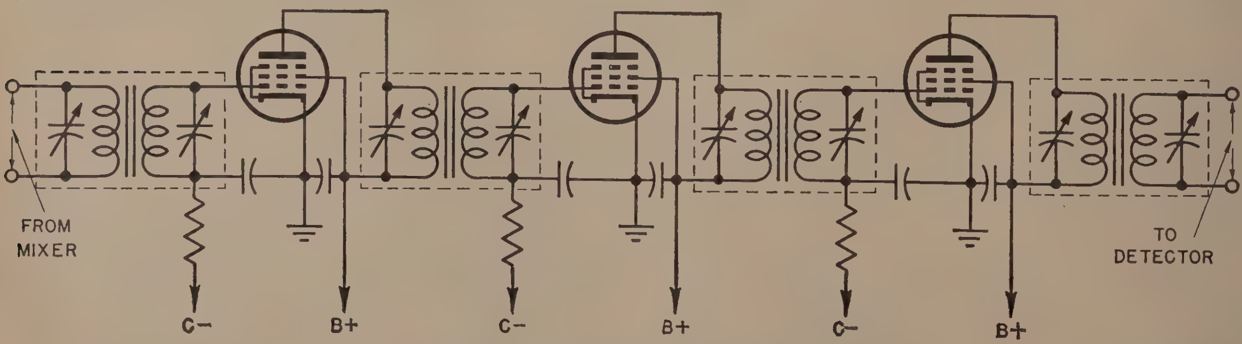


Figure 133. Three-stage i-f amplifier.

73. Tuned Voltage Amplifiers

a. Transformer-Coupled Amplifier. A transformer-coupled amplifier, such as a typical stage in the circuit of figure 133, has a selectivity characteristic that is dependent on the amount of coupling between the primary and secondary of each transformer. The response curves for various degrees of coupling in such circuits are given in figure 134. Since the f-m side bands extend for a considerable distance on either side of the center frequency, a broad flat response is desirable. Beyond this, a sharp decrease in response is necessary to attenuate adjacent channels. An ideal curve for the f-m circuit would be the curve in B, but the sharp corners and vertical sides cannot be attained with conventional tuned circuits. By selecting the proper amount of coupling, such as that of the curve where K, the coefficient of coupling, is equal to .015, an approximation to the ideal curve where K is equal to 1 can be reached in a large number of stages. The value of K is given by the following formula:

$$\text{Coefficient of coupling } K = \frac{1}{\sqrt{Q_p\, Q_s}}$$

where Q_p and Q_s are the values of Q for the primary and secondary, respectively. For example, suppose that the primary and secondary Q's are equal to each other and have a value of 66.7. Then, from the formula,

$$K = \frac{1}{\sqrt{66.7 \times 66.7}} = \frac{1}{66.7} = .015.$$

Such a circuit has the response characteristic indicated by that value of K shown in A of figure 134.

b. F-M I-F Transformers. The operation of the interstage coupling transformer determines the gain and selectivity of the stage. In general, these transformers must be adjustable so that the receiver can be tuned for maximum performance. The transformer can be tuned by variable capacitors across the fixed inductance of the primary and secondary, or the capacitors can be fixed and the inductance of the coils varied by means of powered iron slugs. The coupling of the coils is adjusted at the factory by setting the distance between the primary and the secondary. Low-frequency i-f transformers used in the second conversion section of double superheterodynes are similar.

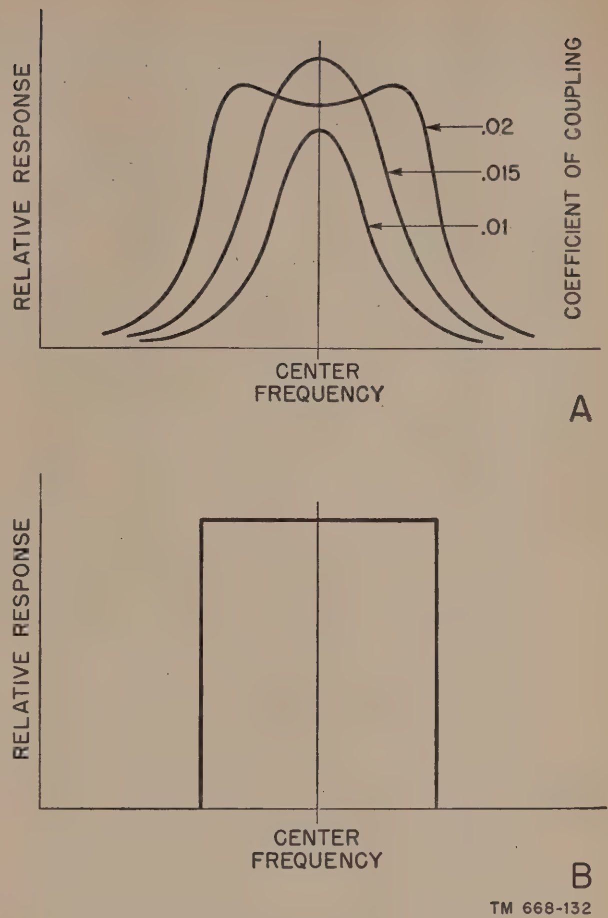


Figure 134. Response of i-f transformers.

c. Distortion in F-M I-F Stages.

(1) To gain the full advantages of f-m reception, the i-f system must have a selective response that does not introduce objectionable distortion into the side-band system. At the same time, it must have sufficient adjacent-channel rejection. If the selectivity curve is not symmetrical about the center frequency, considerable distortion appears in the output of the receiver because some side bands are amplified more than others. The parallel-tuned transformer, in addition to amplitude-response variation, introduces a certain amount of phase shift into the signal. If the phase shift is not uniform with frequency, the effect is somewhat like additional spurious

modulation of the signal. An ideal selectivity characteristic has a uniform phase shift throughout the band of frequencies that it passes.

(2) To correct distortion in practical circuits, it is necessary to make the response curve as flat as possible. Although the double-humped (overcoupled) curve in A of figure 134 provides sufficient bandwidth and steep sides, it has bad phase distortion. However, if a single-tuned circuit is added to this double-humped curve, the shallow depression in the center can be raised, giving a broad flat-topped characteristic with sharply falling sides. Alternate stages are used with either single-tuned circuits or undercoupled transformers and these fill in the depression caused by the overcoupled circuits of other stages. In the typical amplifier, eight sets of coils are contained in four transformers. If two are overcoupled and two critically coupled by the right amount, a curve can be obtained which closely approaches the ideal. Unfortunately, overcoupled circuits are difficult to aline. The tuning of the primary and secondary interact strongly with one another, with the result that the optimum response curve is difficult to obtain without special equipment and tuning procedure.

(3) Another way to achieve the same result is to couple all of the stages critically so that each response curve is essentially that of a single-tuned amplifier. Each stage then is detuned slightly above or below the center frequency (stagger-tuned). The successive stages each provide a portion of the desired flat top of the selectivity curve, and the sides of the curve are obtained by the two stages tuned farthest from the center frequency on either side. No two stages are tuned to exactly the same frequency; therefore, any tendency to instability and self-oscillation is reduced considerably. This method introduces phase distortion into the f-m signal.

d. I-F's for High Adjacent-Channel Rejection.

(1) Where extremely good selectivity is required, the standard double-tuned i-f transformer is not satisfactory. Even in double-conversion receivers there can be enough residual response in the adjacent channel to make reception difficult when a powerful local transmitter is operating while the receiver is tuned to a weak, distant station. Two methods designed to overcome this difficulty are the triple-tuned i-f transformers and the band-pass filter. The triple-tuned transformers have two ordinary windings inductively coupled to each other, but the output of the secondary is fed into a third parallel-resonant circuit through a capacitor. In this way, the coupling of the three circuits can be arranged for good gain and very sharp selectivity. A sufficient number of these stages provides good attenuation of the adjacent channel.

(2) The other method is more complicated, but produces better results. The output of the second mixer is fed into a band-pass filter that contains several tuned circuits. This filter produces a nearly ideal selectivity characteristic. Following the filter is a three-stage resistance-coupled i-f amplifier, which in turn drives the detector. The gain ahead of the filter is deliberately kept low, and the strong adjacent-channel station therefore cannot cross-modulate the weak station to which the receiver is tuned. The filter selects the weak station and rejects the strong one before there is any appreciable voltage gain. All of the voltage gain takes place in the resistance-coupled stages, which receive only the weak signal from the filter, the strong adjacent channel station being almost completely rejected.

74. Stability in F-M I-F Amplifiers

a. Transformer-Coupled Pentode Stage.

(1) If an f-m i-f stage is unstable, the

selectivity characteristic will not be symmetrical, and distortion will be high. Instability may be caused by feedback of energy from the output to the input, either in a single stage or over the chain of amplifiers. This instability usually does not exist on both sides of the pass band of i-f transformers at one time, since the feedback can operate over only a narrow range of frequencies. Therefore, assuming that all tendency to oscillate takes place at or near the center frequency, the amplification will be greater on the side of the response curve that tends toward oscillation. Oscillation may occur in voltage amplifiers at points far removed from the operating frequency, but these are not of immediate concern. An unstable or regenerative amplifier not only has bad selectivity characteristics, but will also respond more strongly to changes in supply voltages, tube characteristics, or signal input. For the best operation of an i-f amplifier, all voltage feedback must be reduced to a minimum, so that interchanging tubes and circuit components for repair does not produce undesirable results. Although it is possible to reduce the tendency to oscillate by reducing the gain of the stage, the loss of gain impairs the sensitivity of the receiver.

(2) The major cause of instability in pentode amplifier stages is the slight residual capacitance that exists between the grid and the plate of the tube, tending to cause tuned-plate, tuned-grid oscillation. The value of this residual capacitance is seldom above .005 $\mu\mu$f, and it exists even if the screen grid is perfectly grounded. This small capacitance is sufficient to promote oscillation when high-gain tubes are used with large values of inductance in the interstage transformers. The feedback can be neutralized as in a conventional triode amplifier, but the small capacitance makes the adjustment of any usual neutralization circuit difficult. The circuit in A of figure 135, shows one method of neutralizing this small capacitance by using a common bypass capacitor for the plate and screen circuit. If properly chosen, this capacitor forms a bridge circuit, as in B, consisting of the grid-plate capacitance, C_{gp}, the grid-screen capacitance, C_{gs}, the plate-suppressor capacitance, C_p, and the neutralizing bypass, C_n. At terminal 1, which is in the grid circuit, no voltage can be received from terminals 2 and 3, the output circuit, because it is canceled out in the bridge. At 10 mc, the neutralizing capacitance is about .002 μf for the typical pentode. The bridge does not

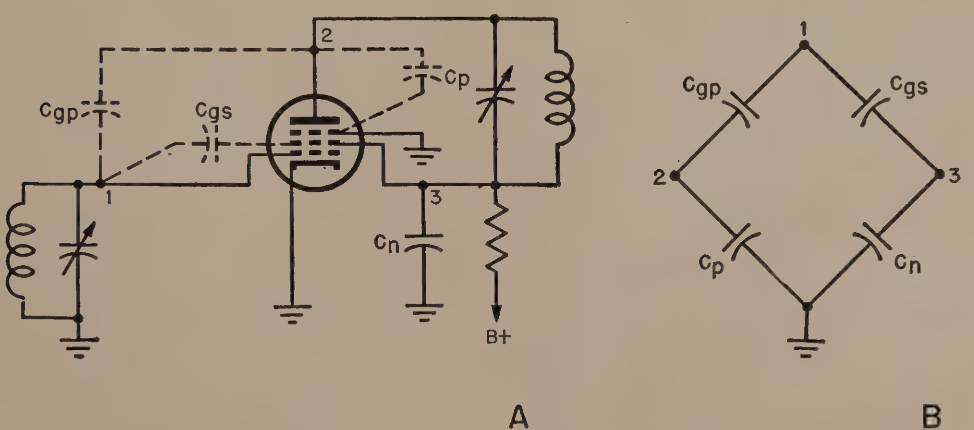


Figure 135. Neutralization of pentode i-f amplifiers.

provide perfect balance, but it materially reduces the tendency toward feedback through the grid-plate capacitance.

(3) Other types of feedback also limit the gain that can be obtained from the stage without oscillation. For example, since the impedance of the circuit is proportional to the inductance and also to the Q, increasing the inductance raises the impedance and the over-all gain. If the inductance is made large enough with the Q held constant, a value is reached where the stage oscillates. Another cause of feedback is the common coupling of the plate and grid circuits through the common current that passes through the cathode. Even with the cathode pin grounded directly, the cathode impedance is far from negligible at high frequencies. This effect is reduced if the tube used has two cathode connections. The plate is returned through its bypass capacitor to the grounded side of the cathode connection, with the grid returned to the ungrounded side, and the paths of the two currents are comparatively independent. When the cathode is above ground, because of impedance in a cathode bypass capacitor, there is a further danger of instability. This is overcome by using capacitors which contain enough self-inductance to make them resonate at the intermediate frequency. In this way they present a very-low impedance path to ground. High-gain i-f stages usually do not use cathode bias.

(4) When the cathode-to-ground path is capacitive, a negative load is placed on the grid circuit, raising its Q. Consequently, its impedance is raised, causing instability. Negative loading can be reduced by using a cathode circuit that is inductive at the operating frequency. Inductive cathode loading places a positive load across the tuned grid circuit which lowers the gain. For further stability, the tuned circuits always are returned to ground at the cathode of each stage. Otherwise, currents that flow through a ground common to the input and the output circuit could interact inductively, causing feedback. To reduce feedback from stage to stage, the grounds for each stage are made at one point, and the over-all grounding plan for the amplifier is arranged carefully to prevent interaction between high- and low-level ground circuits.

(5) Instability is caused between stages by direct coupling through the common filament supply and through a common plate supply impedance. This type of instability is overcome easily with suitable r-f chokes and bypass capacitors placed in the appropriate leads in each stage. Careful layout of the power-supply leads and shielding also is necessary to prevent inductive and capacitive interaction between successive stages. The leads to the transformers and the associated parts within the individual stage must be kept as short as possible to prevent the development of stray capacitance which could cause feedback energy. The use of seamless drawn cans and iron slugs that magnetically shield the ends of the coils reduce the reaction through magnetic and capacitive coupling between stages. Capacitive coupling between the windings of the transformer itself can be reduced by placing an electrostatic grounded shield between the windings, or by using shielded wire for the primary. This prevents transfer of energy by direct capacitance between the coils and confines it to inductive coupling alone.

b. High-Gain I-F Amplifiers.

(1) Since maximum over-all gain in the i-f amplifier permits the detector circuit to operate at the high level desired, some means of increasing the gain of the amplifier must be devised. Increasing the ratio of inductance to capacitance in the i-f transformers increases the gain, but when the capaci-

tance in the grid circuit is reduced to a low value the Miller effect becomes prominent. Miller effect introduces a change in the input capacitance when the transconductance of the tube changes and if very low values of input capacitance are used, the changes are sufficient to detune the stage. When sharp impulse-noise bursts cause the signal at the grid of the last i-f stage to increase, they momentarily swing it far into the cut-off region. This momentary swing causes detuning of the transformer because the input capacitance can change as much as 2 $\mu\mu$f. The sudden change in tuning spoils the symmetry of the amplifier response and produces phase distortion that appears in the output as noise or garbling of the signal. Most of the amplitude variation that is produced by this impulse-noise is removed in the detector, but the phase distortion remains. Since the gain of an amplifier increases as the product of the transformer capacitances decreases, it is possible to overcome the Miller effect on impulse noise by using the stray capacitance to tune the output circuit, and a fairly high capacitance in the input circuit. The use of the stray capacitance to tune the plate circuit of the i-f stage allows the ratio of the inductance to capacitance to be increased Increasing the grid circuit capacitance causes the capacitance resulting from Miller effect to be effectively in parallel with a large capacitance and reduces its effect on the tuning of the stage

(2) A second problem in high-gain i-f systems is the instability that begins to appear when the over-all gain is high enough to supply the 2 volts needed to operate the detector in many receivers. The maximum gain of the i-f stages cannot produce the required output without feedback when the r-f signal at the antenna terminals is small. This disadvantage can be overcome partially with a double-conversion receiver. The second i-f section, however, does not have sufficient bandwidth to handle a wide-band signal, and, because of difficulty with spurious responses from the second oscillator, it is not always desirable to use a double-conversion receiver.

(3) A circuit has been devised for use with both wide- and narrow-band f-m that overcomes this disadvantage. After passing through three high-gain i-f stages, the signal is applied to a frequency doubler. Since an f-m signal is not distorted by doubling its frequency, the character of the modulated wave is changed in no way. However, it does make possible a second high-gain i-f strip that is at twice the frequency of the first. The second i-f operates at a frequency with twice the deviation and bandwidth of the first i-f section. Therefore, the signal applied to the f-m detector has a higher deviation without any significant increase in noise input. There is no danger of interaction between the high-level stage near the detector with the low-level input, since the frequencies are different. With most signals, there is sufficient voltage at the grid of the doubler to make it draw grid current, which makes the output of the doubler substantially independent of the input signal. The result is a limitation of variations in amplitude and consequently improved performance. Limiter stages which follow can have more gain too, because there is less danger of feedback. Finally, there is automatic squelching of background noise if no signal is present at the grid of the doubler, because there is no i-f output at the higher frequency.

75. Limiter Circuits

a. Purpose. A limiter stage generally is an i-f amplifier so arranged that, after a certain point, a further increase in input signal produces no change in the amplitude of the output signal. If there is sufficient gain ahead of such a stage, variations of amplitude in the received signal are removed. Since the detector responds

only to frequency variations, the intelligence contained in the modulation is not impaired; in fact, fading and some types of noise are removed. Limiter stages are used after the last i-f stage in some f-m receivers. The circuit of a limiter is almost identical with that of a standard i-f amplifier, with the exception of the values of the components and operating potentials. Although a limiter provides some amplification before its saturation point is reached, its main purpose is to limit the amplitude variations caused by fading and noise in the output voltage.

b. Operation.

(1) The response of an ideal limiter is shown in A of figure 136. The amplifier tube must have a sharp cut-off characteristic. Very low values of screen voltage are used, and little bias is applied to the control grid. Therefore, large values of positive grid voltage quickly drive the tube to saturation, and large negative values drive it to cut-off. The transfer characteristic of the limiter is shown in B. An input signal varying greatly in amplitude with sharp pulses of noise superimposed is applied to the input of the limiter system. The varying voltage applied at the grid of the tube drives it to cut-off or saturation so that the large variations in amplitude are clipped off. In B, some amplification takes place for voltages between points 1 and 2, but the output is held constant beyond 2. Values more negative than point 1 are below cut-off and therefore do not appear in the output. For proper limiter action, the lowest-amplitude signal must swing at least from 1 to 2. The resultant output signal is square because the extreme positive and negative peaks have been clipped off in the limiter.

(2) The limiter circuit shown in figure 137 is similar to the standard i-f amplifier

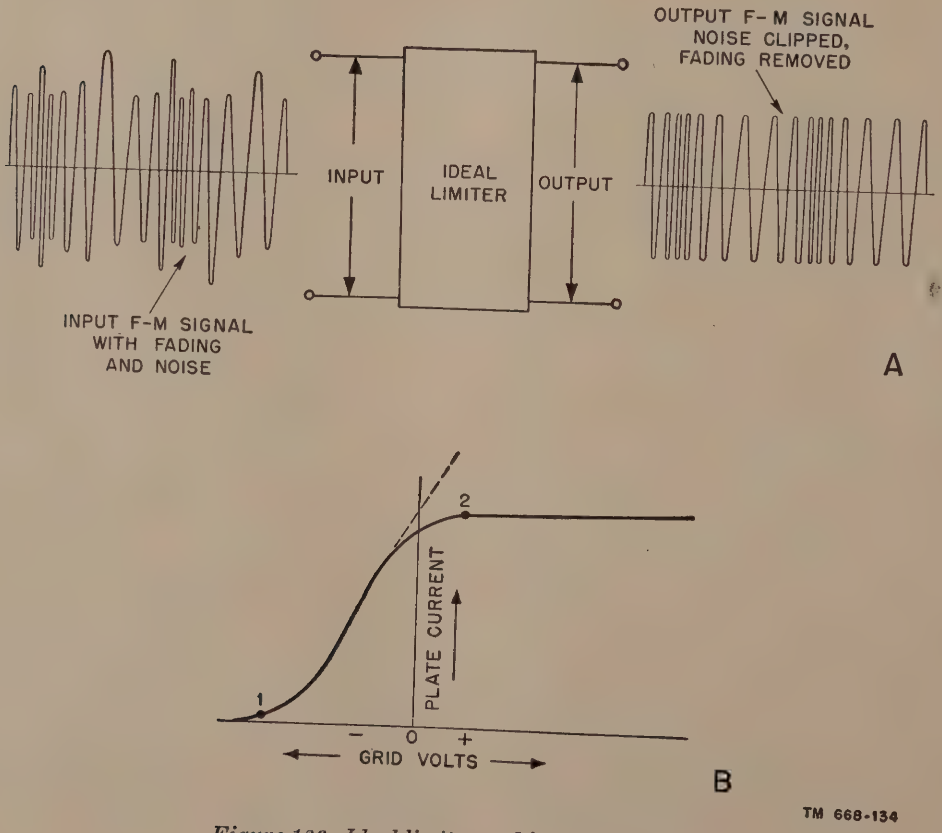


Figure 136. Ideal limiter and its characteristic.

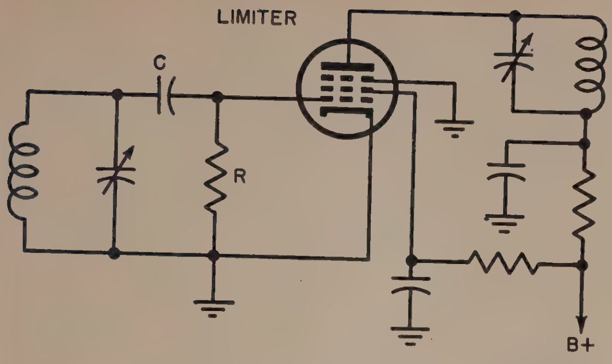


Figure 137. Limiter circuit.

except for the grid-leak resistor and capacitor to provide bias. The R-C combination produces a voltage that is equal to the peak d-c rectified voltage between grid and cathode. The rectification that takes place provides the bias, in addition to its clipping action. The short disturbances in the signal caused by noise voltage that are longer than the time constant of the R-C circuit are clipped off. The longer variations caused by fading of the signal appear in the output. To take care of the slower fading, it is common practice to follow the first short-time-constant limiter with a second limiter that has a longer time constant.

c. Cascade Limiters.

(1) Where it is desirable to eliminate both impulse noise and slow fading in a limiter, two limiter circuits with different time constants are required. The limiters usually are in cascade and

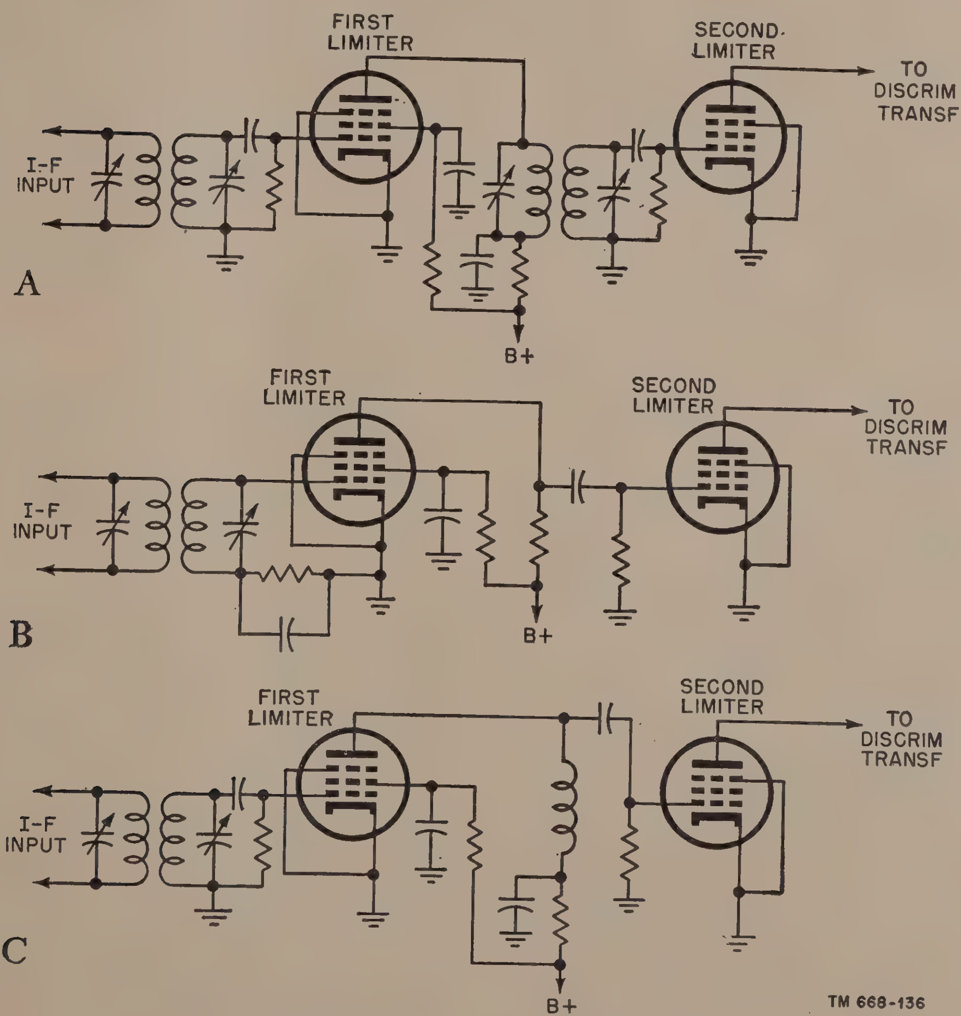


Figure 138. Cascade-limiter coupling circuit.

have different grid-leak values. Several different circuits used to couple such cascade limiters together are shown in figure 138. Transformer coupling, in A, provides greater gain for the second limiter, but is at a disadvantage where there is sufficient i-f gain to saturate the first limiter. Resistance coupling, in B, is most widely used because of its simplicity and ease of adjustment. In C, impedance coupling is a compromise between the other two.

(2) It is not desirable to have any appreciable gain in the limiter because these stages can contribute to the overall amplification of the i-f signal, and may cause regeneration. Amplification in the limiter takes place only when insufficient signal is applied to the grid circuit. Since it takes about 2 volts to saturate the average limiter, if signals of 1 microvolt are to be received at the antenna, a gain of at least 2,000,000 must be effected ahead of the limiter.

d. Miscellaneous Limiters.

(1) Other circuits which serve as limiters, such as saturated diodes and cathode-coupled triodes, are called dynamic limiters. Since the diode ceases to conduct if the plate is made more negative than the cathode, it is possible to clip the negative halves of the signal. If another diode is placed in series with the first, the positive half cycle is clipped also. This circuit produces a square wave from a sine-wave input signal.

(2) A cathode-coupled amplifier also can act as a limiter if proper bias is applied to the grids. As the grid of the first tube goes positive, it draws current. When it goes far negative, the cathode also goes negative, which effectively causes grid current in the grounded grid of the second tube. Output is taken from the plate circuit of the grounded-grid section. The limiting action occurs through grid current on both halves of the cycle. However, since two triodes or a dual triode are required, with little savings in parts, the circuit is infrequently used.

Section VII. F-M DETECTORS

76. Double-Tuned Discriminator

The double-tuned discriminator described in chapter 4 was discussed in conjunction with frequency-control circuits. In f-m receivers the discriminator converts the frequency variation of the signal into audio variations. The transfer characteristic of the discriminator (fig. 139) is plotted in respect to the change in frequency and the amplitude of the audio output voltage. As the frequency of the input varies, each frequency deviation is translated into an effective change in amplitude of voltage at the discriminator output. When the frequency deviation reaches its peak value on either side of the center frequency, the audio voltage also reaches a peak value. Changes in the rate of frequency deviation produce variations in the rate of change of audio voltage which are equivalent to changes in its frequency. The double-tuned circuit is used infrequently in f-m receivers because the three tuned circuits required in the transformer are difficult to aline, and the design of the transformers becomes more critical.

77. Phase Discriminator

a. The phase discriminator, previously discussed in relation to frequency control circuits, is one of the most frequently used f-m detectors. The principal advantage of the phase discriminator over the double-tuned discriminator is its ease of alinement. Since the discriminator has no inherent rejection of amplitude modulation, it always is operated after a limiter stage. The secondary inductance of the discriminator transformer is lower than that of the primary to provide a step-down from the high plate impedance of the limiter to the low input impedance of the diodes. Therefore, the transformer for a phase discriminator differs somewhat from

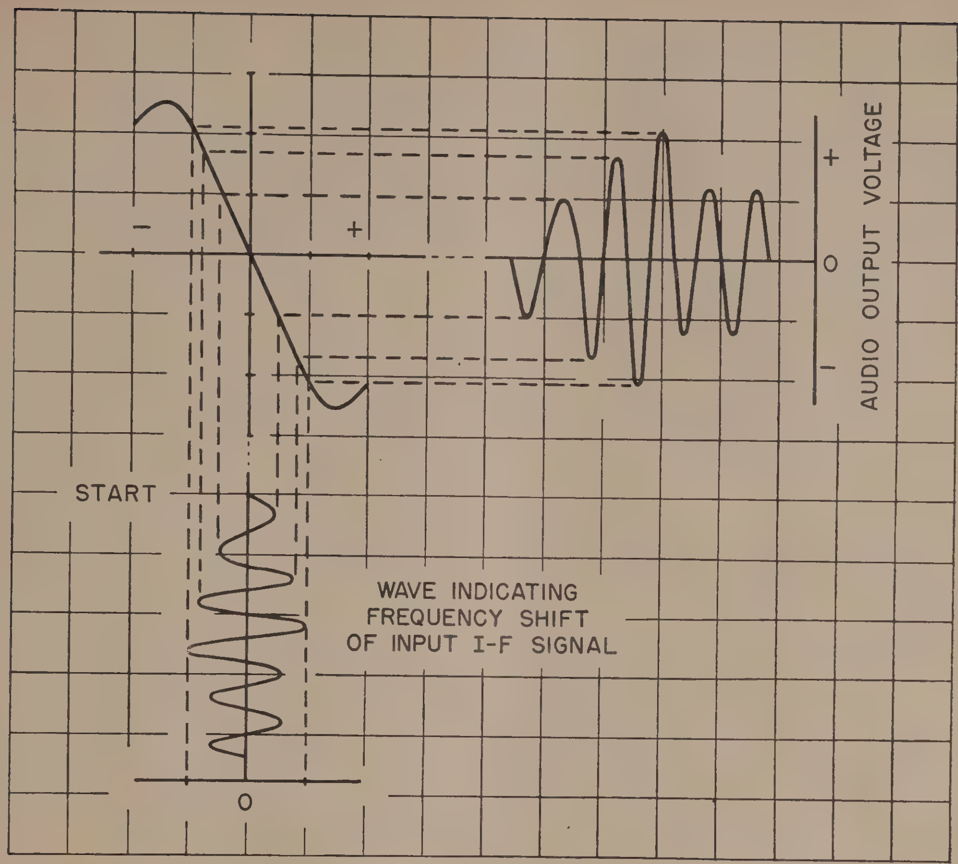


Figure 139. Transfer characteristic of double-tuned discriminator.

that used between two i-f stages. Diodes present a much lower impedance to the load than do the grids of a voltage amplifier, and since they vary from conducting to nonconducting conditions during various parts of the cycle, the actual load presented to the tuned circuit also changes drastically.

b. The modified phase discriminator has essentially the same properties for f-m reception as the ordinary phase discriminator, but fewer components are needed. It is possible to recover recognizable audio signals at frequencies other than the center of the discriminator frequency with any type of discriminator because the selectivity curve of the transformers continues beyond the characteristic **S**-shaped transfer curve. On the sides of these curves a signal that varies with frequency can be rectified by the diodes. When the receiver is slightly detuned, an audible response is present, although with considerable distortion. This occurs on both sides of the i-f channel, and detection off the center frequency is called side response. With a discriminator, the side responses are sometimes only 20 db lower than the desired channel, which shows the necessity for high adjacent-channel selectivity in the i-f amplifier.

c. A modification of a phase discriminator, designed for operation with a tube that has two diode plates, but only a single cathode, is shown in A of figure 140. The circuit differs from the previous ones in the way the reference voltage is applied to both of the diode plates. Separate windings for each diode are connected by the small capacitor, *C*4. Both windings are tuned by capacitor *C*3, so that the effective resonant circuit is similar to that in other phase discriminators. The i-f voltages are applied across each diode as before, with resistor *R*1 acting as the load for the upper diode; *R*2 acts as the load

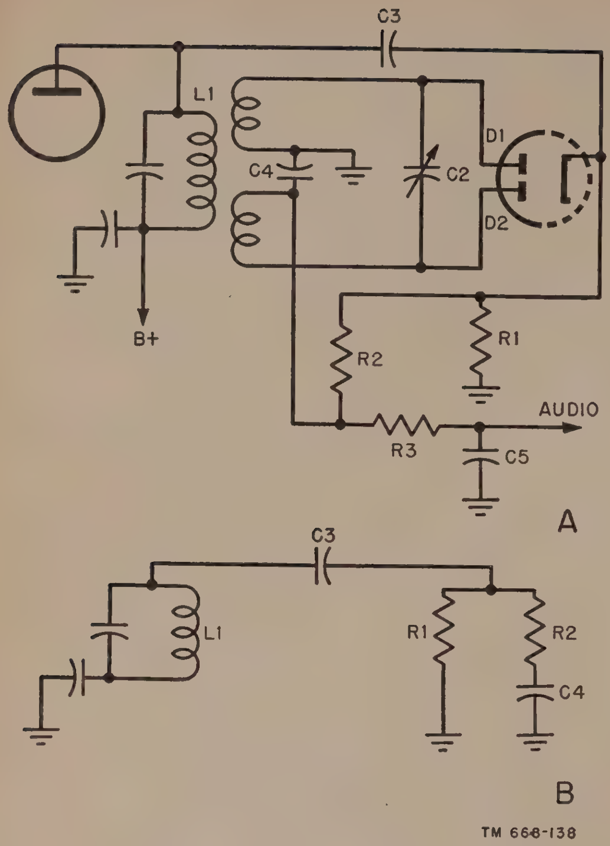


Figure 140. Modified phase discriminator.

for the lower diode because it is returned to the center of the coil and the d-c path is directly from cathode to plate. Although the resistors are connected to a common cathode, they serve as separate loads.

d. The reference voltage is coupled to the cathode by capacitor *C*3, and capacitor *C*4 serves as an r-f bypass to ground for the lower winding. From the i-f stage, there are parallel paths from the plate of the tube, shown in simplified form in B. From the top of the primary, the i-f current passes through *C*3, and then flows through two paths, one through *R*1 to ground, the other through *R*2 and *C*4 to ground. Since the reactance of *C*3 and *C*4 is negligible at the i-f frequency, all of the reference voltage appears across *R*1 and *R*2, as desired. The audio output appears on the high-potential side of the load resistors, which coincides with the lower end of *R*2. The additional resistor and capacitor, *R*3–*C*5, form a de-emphasis network.

78. Ratio Detector

a. Basic Ratio Detector.

(1) The basic purpose of a discriminator is to rectify two i-f voltages whose amplitudes depend directly on frequency. These rectified voltages then are combined so that no voltage appears across their output at the center frequency of the i-f amplifier. A difference voltage proportional to the difference in frequency of the two applied i-f voltages is produced when the frequency of the i-f signal is above or below the center frequency. This detector is insensitive to changes in amplitude at the center frequency, but changes in amplitude off center may cause the audio output to vary. Therefore, whenever a discriminator is used, it must be preceded by a limiter that requires more circuits and tubes. This disadvantage can be overcome by a *ratio detector* circuit which splits the rectified voltages in such a way that their *ratio* is directly proportional to the ratio of the applied i-f voltages, which vary with frequency.

(2) When the sum of the rectified voltages from the transformer is maintained at a constant value, the ratio between them must remain constant, and the individual rectified voltages also must be constant. Output, therefore, is independent of amplitude variations in the signal and no limiter is necessary. A simplified ratio-detector circuit (A of fig. 141) shows both diodes connected so that their output adds, instead of subtracting as in the discriminator. Capacitors C_L across the load resistors have a large value of capacitance and are charged by the output voltage of the rectifiers. This tends to make the total voltage across the load constant over the period of the time constant, $R_L C_L$, since a large capacitor across the combined loads maintains an average signal amplitude that is adjusted automatically to the required operating level. The rectified output

must not vary at audio frequency, and the time constant of the capacitor and the load resistors must be great enough to smooth out such changes. This time constant is approximately 2/10 second. The basic phase comparison circuit and the appropriate vector diagram of the ratio detector and the phase discriminator are the same.

b. Practical Ratio Detector.

(1) In the circuit for a practical ratio detector (B of fig. 141) the voltages, $E1$, $E2$, and $E3$, are obtained in the same way as in the modified phase discriminator. Therefore, the applied voltage to the diodes also is the same. The diodes are connected in series, and the current through load resistor R_L is always in the same direction. Consequently, R_L acquires the polarity shown when the current flows from the plate of $D1$ to the cathode of $D2$. When an unmodulated signal is applied to the primary of the transformer, equal and opposite voltages $E2$ and $E3$ are developed across the secondary in respect to the center tap. These voltages are rectified by the diodes, with the output voltage across the load resistor, equal to their sum, or $E2$ plus $E3$, and the large capacitor, C_L, is charged to this constant voltage. The time constant of R_L and C_L is long compared with the lowest audio frequency.

(2) Since the voltage across C_L is constant, the sum of the voltages across $C3$ and

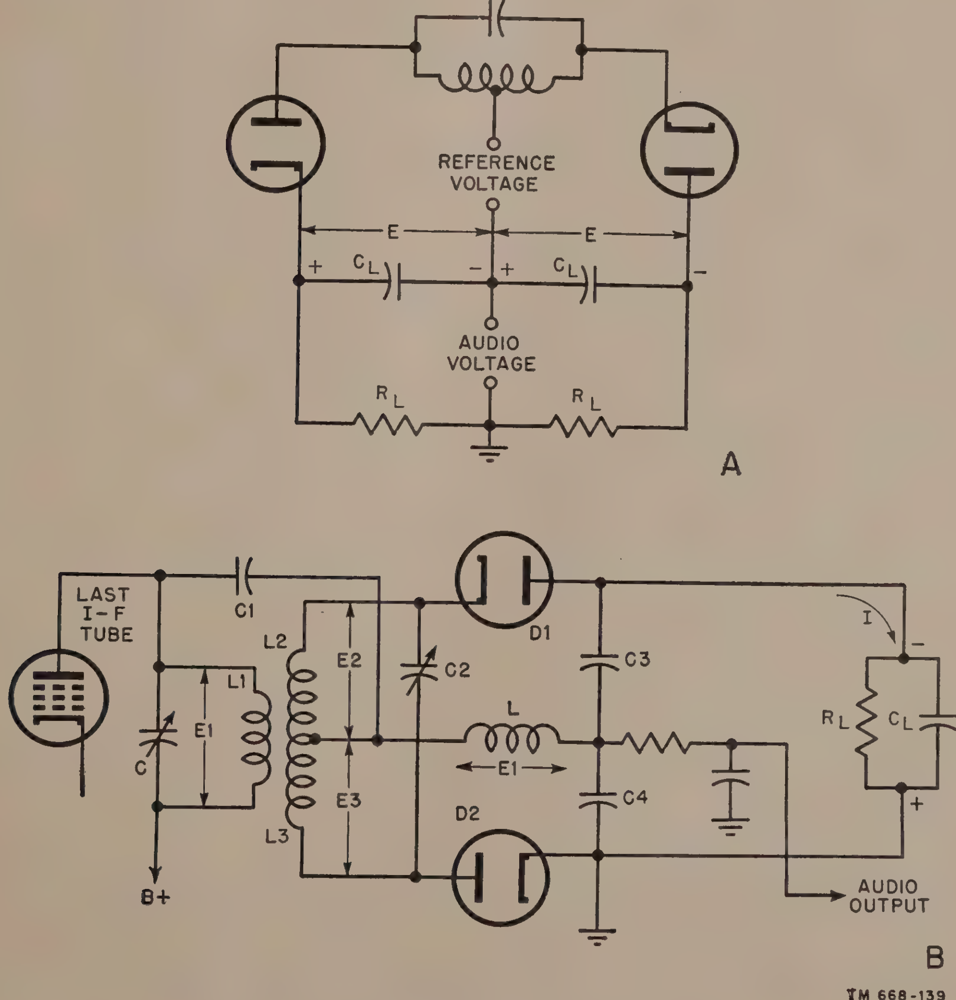


Figure 141. Ratio detector.

$C4$ also must remain fixed. When the carrier frequency shifts with modulation, however, the voltages across $C3$ and $C4$ change, but the sum of their voltages stays fixed at the amplitude of the charge on C_L. When the frequency decreases, $C4$ acquires a greater charge than $C3$; when the frequency increases, $C4$ loses charge to $C3$. Therefore, the voltage between the center tap of the two capacitors and ground varies as the ratio of the voltages across $C3$ and $C4$, the ratio depending on the instantaneous frequency. A variable voltage whose amplitude depends on the frequency deviation of the carrier consequently can be applied to the audio output. As the rate of variation increases with frequency deviation, the voltage at the center tap changes frequency, producing a higher audio frequency. Any amplitude variation in the input signal to the transformers, no matter where the carrier is in its swing, also tends to change the voltage across $C3$ and $C4$. The voltage across the R-C network, however, cannot change rapidly enough to follow the amplitude modulations, and the ratio of the voltage across $C3$ and $C4$ do not change enough to produce an audio output.

c. Performance of Ratio Detector.

(1) The rectified voltage across the load circuit of the ratio detector adjusts itself to the amplitude of the input signal, and there is no minimum level where amplitude variation still can appear in the output. No matter how weak the signal is, the amplitude variations are removed to some extent by the constant charge on the capacitor. However, if signals of greater strength are tuned in, the charge on the capacitor is increased, and the total voltage across $C3$ and $C4$ is increased. Consequently, ratio detectors produce audio output that is proportional to the average strength of the received signal. Ratio detectors can operate with as little as 100 millivolts of input, which is much lower than that required for limiter saturation, and less i-f gain consequently is required. This receiver also is relatively quiet when no signal is received, since tube noise is not amplified as much.

(2) As shown by the curves in figure 142, the tuning characteristic of the ratio detector has much lower side responses than the discriminator because they contain appreciable amplitude modulation which is rejected in the load circuit. The disadvantages of the ratio detector are its greater susceptibility to impulse noise and fading, greater

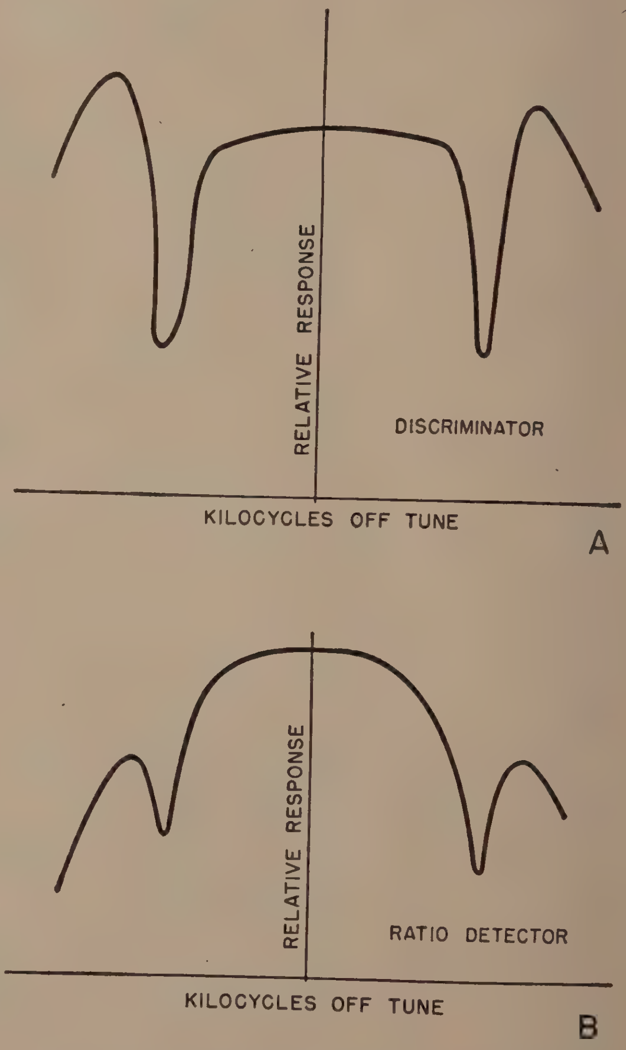


Figure 142. Tuning characteristics of discriminator and ratio detector.

difficulty in alinement, and more complicated transformer design.

d. Modified Ratio-Detector Circuits.

(1) Two modifications of the ratio detector circuit are shown in figure 143. The modification in A allows simpler construction of the transformer. It consists of the addition of a third winding, *L*, from which the reference voltage and the audio output are derived. This tertiary winding consists of a few turns closely coupled to the lower end of the primary. It is essentially a low-impedance source since it is untuned, and the voltage induced in it from the primary is 180° out of phase with $E1$. Voltages $E2$ and $E3$ are each 90° out of phase with $E1$, and the same amount out of phase with the reference voltage which appears across *L*. Capacitor $C5$ is an r-f bypass effectively grounding *L* at the intermediate frequency. Capacitors $C2$ and $C3$ function as ratio capacitors.

(2) The net diode current flow through *L* and $C5$ is zero when the diode currents are equal, because the currents flow through *L* and $C5$ in opposite directions when the carrier is at the resonant frequency of the secondary. When the frequency shifts, the diode currents are unbalanced, producing an audio voltage across $C5$, because it has a high reactance at audio frequencies. This modification overcomes the difficulties of a low-impedance source requirement for the reference voltage, $E1$. At the same time, since it permits the use of a high-impedance primary in the plate circuit of the last i-f transformer, high gain is obtained.

(3) The ratio detector in B has no ratio capacitors, and the cathode of the lower diode has been grounded. The voltage across the tertiary winding is applied to both of the diodes to produce a reference voltage. The i-f current now passes through the large capacitor, $C4$, since the reactance is low and it does not impede the flow of the i-f current. The operation of this circuit is similar to that of the modification shown in A.

79. Synchronized-Oscillator Detector

a. Frequency-Dividing Locked Oscillator.

(1) Using a pentagrid converter tube, an oscillator can be constructed which operates at a frequency that is a fraction of the receiver i-f. If the output of the i-f is applied to what is normally the oscillator grid, with oscillations taking place at the signal grid, the oscillating converter tube will follow the frequency deviation of the i-f at a fraction of the frequency. The oscillator at a lower frequency follows the deviation of the i-f, and the deviation of the locked oscillator is reduced. This reduced deviation can be detected in a conventional discriminator to recover the audio signal.

(2) The oscillator synchronizes with a voltage that is 1/20 of the output oscillation amplitude. This provides an effective voltage gain of 20 on a lower frequency which does not endanger the stability of the i-f amplifier. Moreover, the input to the oscillator has no effect on the amplitude of oscillation. The output is like a limiter in that amplitude variations are removed. Since the discriminator operates on a much lower frequency, the advantages of the high selectivity obtained with double conversion are realized without additional i-f amplifiers or converter circuits. This circuit also discriminates to some extent against impulse noise. Noise immunity is about that of a cascade limiter.

(3) The disadvantages of this system result from the imperfect tuning and insufficient signal strength. Unless the receiver is perfectly tuned, the oscillator does not follow the extreme deviation limits, and distortion is produced in the output. When the signal strength is not sufficient to synchronize the oscillator, no recognizable output at all can be obtained. A more

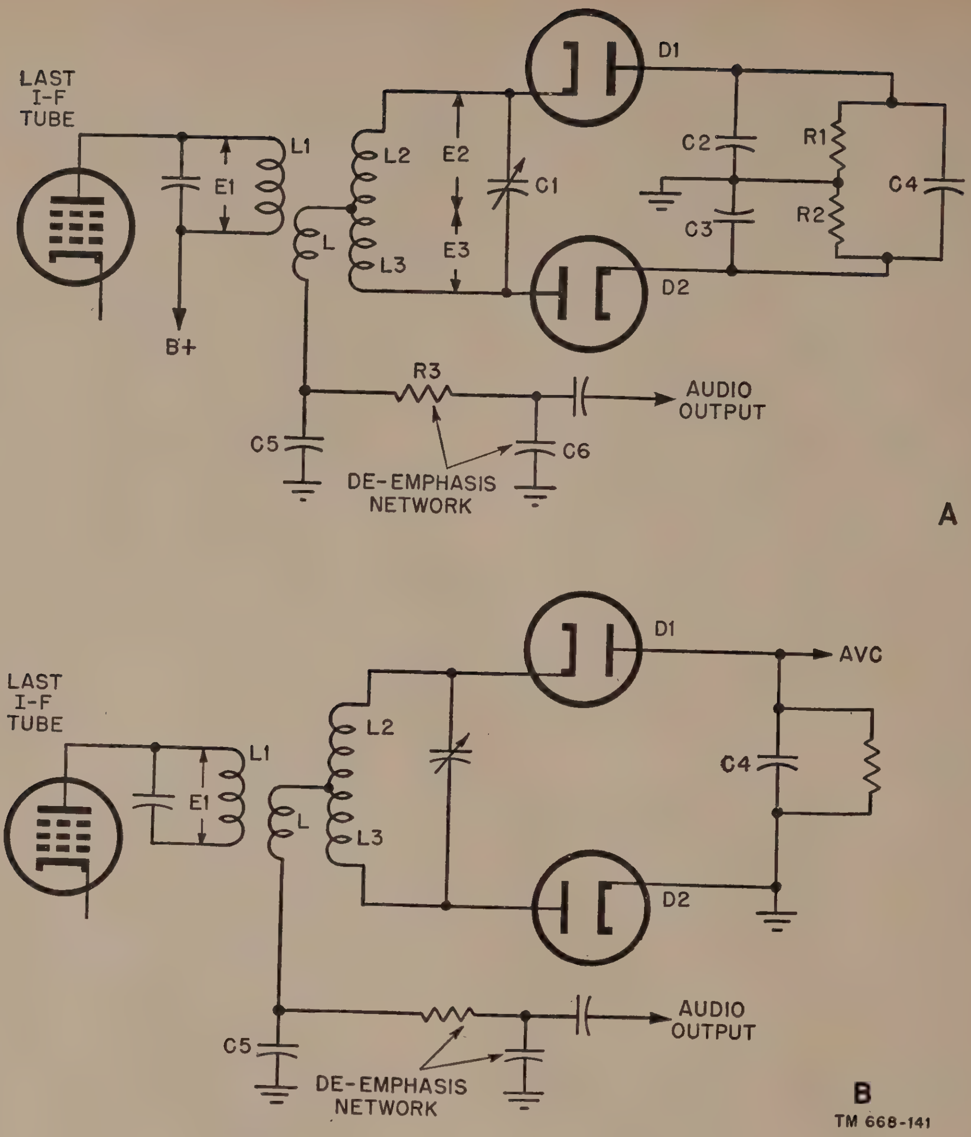


Figure 143. Two modifications of ratio detector.

complicated modification that involves a reactance modulator to keep the oscillator synchronized also has been tried, but the complexity of the resulting circuit makes it impractical.

(4) A major advantage of the frequency dividing locked-oscillator circuit is its rejection of undesired signals in the desired channel. In a conventional f-m receiver, if two stations are on the same frequency but are different in amplitude by 3 db, the stronger signal appears about 9 db greater in the output. With a locked-oscillator detector, a signal that is 3 db stronger than another on the same channel can capture the oscillator almost completely, reducing the level of interference by more than 30 db. This is a great improvement over the other types of detectors.

b. Single-Stage Locked Oscillator.

(1) The frequency-dividing locked oscillator is really not a detector since the actual detection is carried out in a modified discriminator circuit. A special circuit using a tube that incorpo-

rates the detector and the oscillator in a single envelope is illustrated in A of figure 144. The cathode and the first two grids constitute the tube for the locked oscillator, which is a Colpitts circuit. The signal at the oscillator frequency is coupled to the plate of the tube through the electron stream. The oscillator tank circuit is formed by the variable capacitor, $L2$, C_a, and C_b. Grid-leak bias is provided by $C4$ and $R1$. The second grid acts as the anode of the oscillator section and is at ground potential for i-f, since it is bypassed by capacitor C_c. The fourth grid is connected internally to the second grid and acts as a shield for the signal grid, number 3, since it is also at ground potential.

(2) The plate circuit is formed by $L3$ and $C3$ in parallel. All of the tuned circuits are resonant to the i-f. The Q of the plate tank circuit is lowered by resistor $R3$ in parallel with the coil and therefore its bandwidth is increased. The impedance of this circuit does not change appreciably throughout the i-f pass band. With no signal applied to grid 3, oscillator pulses flow through the plate tank circuit and develop a voltage across it. When a voltage is applied to the signal grid, the amount of plate-current flow changes, and the average value of the pulses of current appearing in the plate load also varies.

(3) A vector diagram of the operation of the stage is shown in figure 144. Vector e_1 is the oscillator voltage on grid 1. Vector e_{3a} is the voltage on grid 3 when the incoming frequency is the center of the i-f band. When there is no signal input, or when the signal is unmodulated, the pulses of current flowing in the tube do not change.

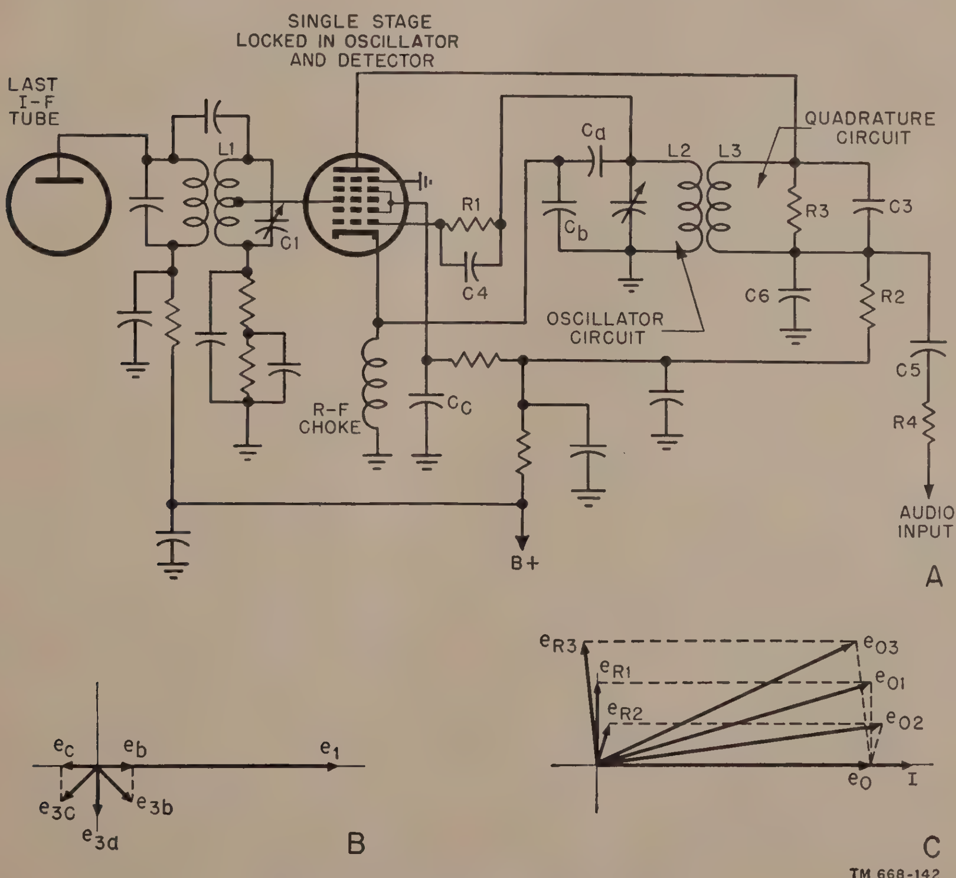


Figure 144. Locked-oscillator detector.

When the incoming signal and the pulses differ in phase, a component of the input signal is either in phase or out of phase with the pulses. The amplitude of the pulse voltage increases or decreases. This phase change is produced when the input voltage frequency varies. Vector e_{3b} shows the signal voltage when the frequency is below the center, and e_{3c} when it is above. These variations produce changes in the main vector, e_1, as shown by the horizontal components, e_b and e_c, that add or subtract from it. The magnitude of the current flowing in the plate circuit therefore varies with the frequency of the incoming signal. The average value of the pulses is obtained in the filter circuit of resistors and capacitors connected to the audio output.

(4) The frequency of the oscillator circuit is changed by feedback coupled from the output circuit. Feedback voltage is proportional to the amplitude of current pulses and is 90° out of phase with oscillator voltage because of the action of the tuned circuit. The circuit acts like an inductively coupled reactance modulator which varies the effective inductance in the oscillator tank circuit. The oscillator therefore tends to stay locked in with the frequency of the incoming signal, as in C. Vector e_O is the voltage across the oscillator tank without feedback. The current flowing through the tank is in phase with the applied voltage, as shown by the current vector, I.

(5) With feedback, the voltage across the tank is the voltage induced by coupling and the vector, e_O. The induced voltage is 90° out of phase with e_O, and this is equivalent to introducing inductance in series with the tank circuit coil. This increased inductance synchronizes the oscillator with the frequency of the incoming signal. With the oscillator exactly synchronized and with no modulation, a voltage is introduced in the oscillator circuit, e_{R1}. With e_O, this produces the resultant vector, e_{O1}. When the incoming frequency increases, the pulses of plate current decrease, and the voltage induced in the oscillator tank also decreases. This is shown by the smaller vector, e_{R2}. The effective inductance is decreased, and the frequency of the oscillator is increased until it is brought back into synchronism with the input signal. When the frequency of the input signal decreases, the reverse operation takes place, as shown by vector e_{R3} and resultant oscillator voltage vector e_{O3}. The plate current changes linearly with these frequency variations. At the same time, it holds the oscillator in synchronism with the applied frequency. The plate current variations are recovered through the audio load resistor, *R*2. Capacitor *C*6 serves to bypass i-f current. The audio output is coupled to the following stages through *C*5 and *R*4.

(6) This circuit effectively suppresses amplitude modulation in the same way that the locked oscillator divider does. In addition, it acts automatically as its own detector. Like all locked oscillator detectors, it has a threshold amplitude of input signal below which no signal at all is received. The input signal required to lock the oscillator is approximately 1 volt, which is slightly less than that needed for a limiter. Amplitude-modulation disturbances appear in the output as distortion because the oscillator is forced out of synchronism for short periods of time, with resultant loss of linearity in plate-current change. This circuit combines the function of limiter and detector in one circuit, with good noise immunity. The audio output is independent of the input signal, although the distortion decreases with increasing signal input. The receiver is quiet in the absence of a received carrier because there is no direct connection between the detector circuit and the i-f amplifier, provided the shielding is adequate.

80. Cycle-Counting Detector

a. An f-m detector that requires no alinement at all, and which operates with a resistance-coupled second i-f amplifier, is shown in figure 145. Selectivity is obtained through the use of a tuned first i-f at high frequency. The resistance-coupled circuit will pass a band of frequencies from low audio frequencies up to about 200 kc. The essential operating feature of this detector is its response to the number of cycles provided by the two-stage resistance-coupled limiter. These limiters produce square waves which are rectified by the dual-diode, whose output consists of negative-going pulses. The positive halves of the square waves are shorted to ground by the first half of the diode. The second half of the diode passes the negative halves, which charge the capacitor connected from plate to ground. As the frequency increases, the average rate of charge increases, and the voltage across the capacitor increases. Starting at a given center frequency, the charge on the capacitor fluctuates, depending on the departure of the signal from that center frequency. The resistor serves to lower the time constant so that the storage of charge can change fast enough to reproduce an audio signal.

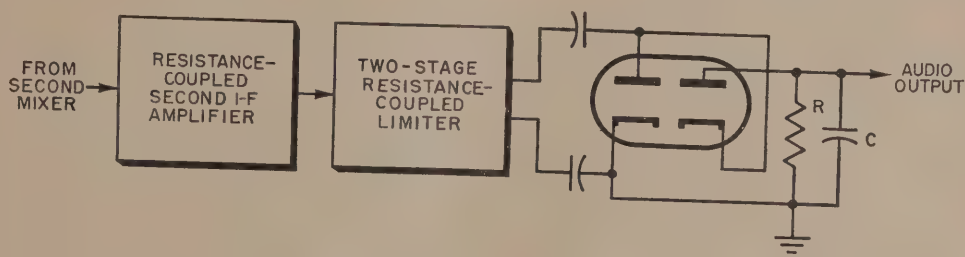


Figure 145. Cycle-counting detector.

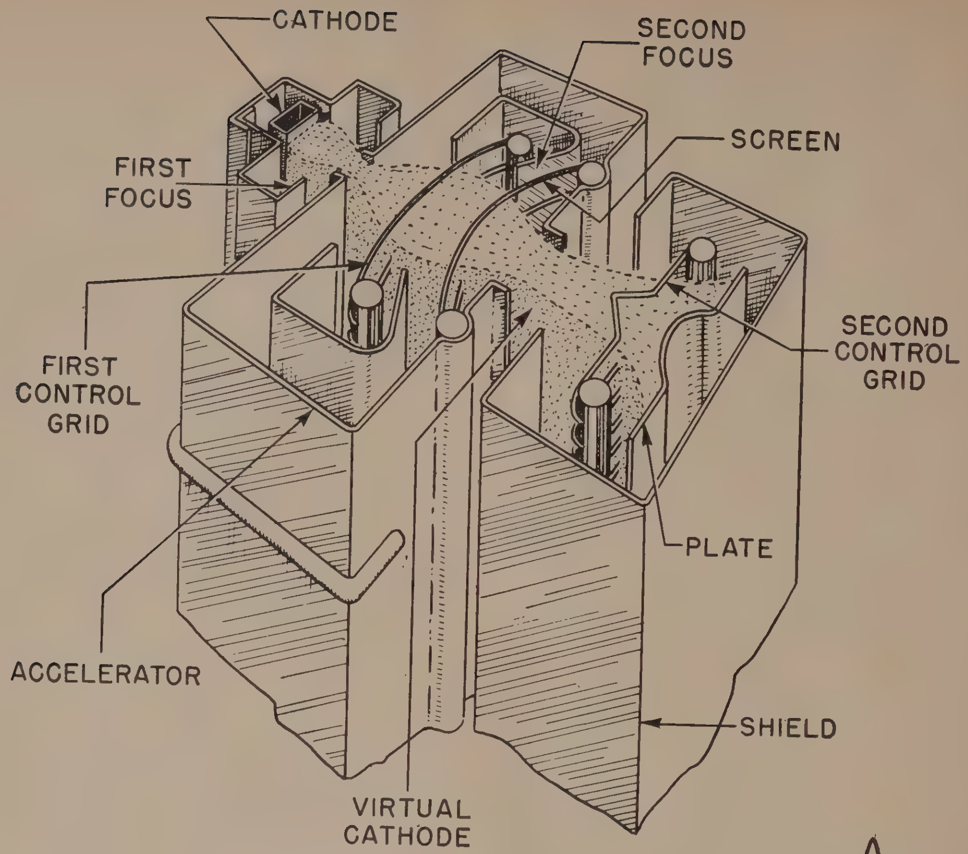
b. The dual limiters reduce undesirable a-m, and the detector responds only to frequency-modulated signals. The output is coupled through a capacitor to the first audio stage. Besides the ease of alinement, this circuit has fairly good sensitivity because of the dual i-f system. The selectivity, however, tends to be poor because of the small number of tuned circuits in the high-frequency i-f. The circuit has low distortion and is used widely in frequency monitors for f-m transmitters. The detector produces output as low as 0 cycle (d-c), which can be used to actuate meters that show the departure of the master oscillator from the assigned frequency.

81. Gated-Beam-Tube Detector

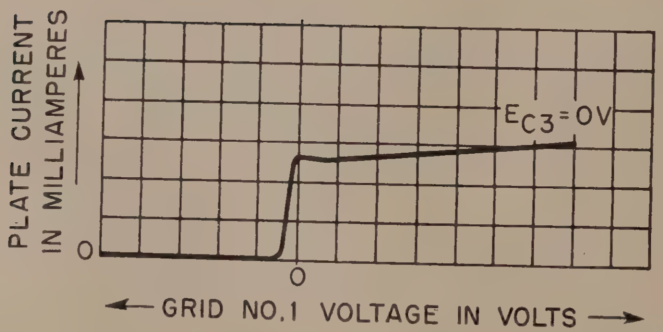
a. Tube.

(1) The gated-beam tube (fig. 146) is a highly efficient limiter and a special type of discriminator contained within one envelope. Its operation depends on the behavior of focused streams of electrons, and, although it has grids and a plate, its operation is different from that of most receiving tubes. With the construction shown in A an unusually sharp cut-off tube results and the transition between anode current flow and cut-off, as controlled by grid voltage, is very abrupt. At the point where current begins to flow, the transconductance is much higher than for any other type of tube. This results in the step-shaped control characteristic shown in B, where plate current is plotted against grid 1 voltage. After the plate current rises to its maximum value, no change in grid voltage can produce any further change in plate current. *This control characteristic is found only in this tube.*

(2) The electron stream from the cathode is passed through focusing and accelerating electrodes which form an electron beam. The beam then passes through the first control grid if this grid is either zero or positive. The electrons continue through a second focusing electrode, a screen grid, and then through a narrow slit which acts as a virtual cathode for the second



A



B

TM 668-144

Figure 146. Gated-beam tube.

control grid. The control characteristic of the second control grid is similar to that of the first grid and the screen grid is inserted between them to act as a shield. Electrons pass through the tube in a flat sheet which narrows down at the focus and accelerator electrodes and broadens out in the vicinity of the control grids. When the potential is close to zero, the electrons in the vicinity of the control grid move slowly, although most of them travel in substantially straight lines when passing through it.

(3) If the voltage on the first control grid is made a trifle more negative, a few electrons are repelled in front of it, rapidly building up a space charge

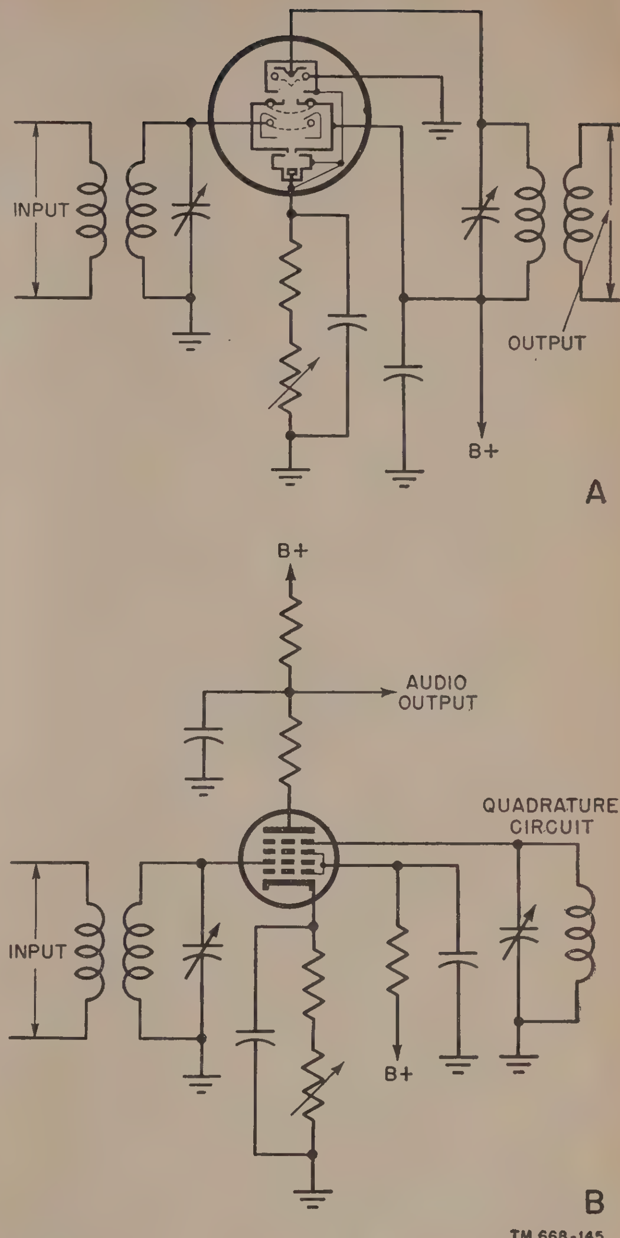
which is concentrated in the center of the beam, forcing the remaining electrons to the side. Those falling back miss the small narrow opening from which they emerged, and the result is an abrupt cut-off of plate current, as shown in B. The operation of the second grid is identical with that of the first except that, since fewer electrons are available, the transconductance is slightly less.

(4) When the control grids are driven positive, they draw current. However, they cannot draw more than a proportionate share of the total beam current. Those electrons not captured by the positive control grid are accelerated and pass through its openings. If the accelerator and anode voltages are sufficiently high, the electrons are moving so rapidly when they near the positive grid that it captures few of them. The result is that the grid current is low, leveling off at a maximum value of about 500 microamperes long before maximum positive grid potential is reached. The entire complex assembly is contained within a seven-pin miniature-tube envelope.

b. Operation as Limiter.

(1) The circuit for the gated-beam tube operated as a limiter (A of fig. 147) is much like that of an ordinary amplifier, but its limiting capabilities are excellent because of the tube-control characteristic. Cathode bias is used, giving an effective negative first control-grid potential of about 1 volt. This corresponds to the center of the steepest part of the control characteristic curve. Since the bias is critical, a small variable resistor is placed in series with the cathode so that field adjustments can be made. The control grid is returned to d-c ground through the secondary of the input transformer. The second control grid usually is grounded and plays no part in the operation of the tube as a limiter.

(2) Limiting occurs instantaneously if the voltage on control grid 1 is shifted slightly in either positive or negative direction. The application of signal voltage turns the tube plate current on and off with each cycle like a switch. The plate current then consists of positive-going pulses which are flat-topped because the plate current levels off with the application of more than 1 or



A. Limiter circuit using gated-beam tube.

B. Limiter-discriminator, audio-amplifier circuit using gated-beam tube.

Figure 147. Circuits.

2 volts of positive grid signal. Since there are no resistors or capacitors in the grid circuit, this limiter has a short time constant, and impulse noise, which is shorter in duration than a cycle of signal voltage, is clipped almost instantaneously. Consequently, this tube, as a limiter, produces better noise immunity than any other f-m limiter. Operation is extremely simple and stable. Only one stage is needed with far fewer parts than are required for a cascade limiter, whose performance it exceeds.

c. Operation as Discriminator.

(1) When the first control grid is biased at the base of its control characteristic, it passes the beam during positive half-cycles and rejects it during negative half-cycles. The square electron pulses pass through the second accelerator and form a space charge which varies periodically in front of the second control grid. By this space-charge coupling, a periodic charging current of about 15 microamperes per megacycle is produced in the ground-return circuit of the second control grid. If a tuned circuit is inserted in this grid lead, as in B, about 5 volts are built up which lag the input voltage on the first grid by 90°, when the circuit is at resonance. This quadrature voltage operates the second grid on the steep portion of its control characteristic in the same way as the first grid. The result is that a plate current flows which consists of pulses passed by the gating action of the two grids. The beam can reach the plate only when both gates are open, plate current flow starting with the delayed opening of the second grid and ending with the closing of the first.

(2) When the signal to the first grid is frequency-modulated, there is an instantaneous shift in phase between the two grids. The Q of the tuned circuit tends to maintain the frequency, and gating action remains relatively constant on the second grid. The first grid, however, changes in phase as the frequency varies. This of course changes the timing of the opening of the two grids and varies the length of time during which plate current can flow. Consequently, the average plate current varies with the modulating frequency. The plate current pulses charge a capacitor from plate to ground through a small resistance. The audio voltage appears across the capacitor. The typical response curve has a linear relation between deviation and audio output voltage over the entire range of frequencies at which the Q of the quadrature circuit holds up its voltage. This means that a receiver using this tube has a wide range of correct tuning within the i-f pass band and produces little signal outside of this range. The result is better adjacent-channel selectivity than can be obtained with a discriminator or ratio detector.

(3) The performance of the circuit depends on the Q of the quadrature circuit. Higher Q's give greater audio output with a consequent loss of bandwidth. A small resistance in series with the plate circuit causes i-f voltage to appear at the plate. This i-f voltage is coupled through the interelectrode capacitance between plate and second control grid into the quadrature circuit. It is in phase with the quadrature voltage and therefore it aids in driving the tuned circuit. The result is a higher quadrature voltage and greater output.

(4) Effectively, a very linear discriminator and an excellent limiter are combined in one tube, with few parts and with noncritical adjustments. After the limiting characteristic is adjusted by setting the cathode-bias resistor, it is necessary only to tune the quadrature circuit for maximum audio output. High-frequency operation is limited by the capacitance between the signal grid and the quadrature grid. The capaci-

tance is reduced greatly by the screen, but it is sufficient to permit a degenerative, out-of-phase voltage to appear across the quadrature-tuned circuit at high frequencies, thereby reducing its voltage. With great care in shielding, operation can be obtained up to about 30 mc. The sensitivity and a-m rejection decrease as the i-f is increased. Since about 1 volt is needed to operate the tube with full limiting, the same amount of i-f gain is needed as for the pentode grid-bias limiter. However, since the beam tube does not have any gain at the i-f, there is much less danger of instability.

Section VIII. SQUELCH CIRCUITS

82. General

Squelch circuits remove the background noise present in limiter-discriminator detectors when no signal is present. In a high-gain receiver, the noise output is sufficient to be very annoying to operators who must monitor a channel for long periods of time. The squelch circuit lowers the audio output of the receiver when no signal is being received, and allows it to operate normally when a signal is present.

83. Limiter-Derived Squelch

a. Direct-Coupled Cascade Limiter.

(1) The limiter circuit in A of figure 148 produces a partial squelch action. The signal from the i-f amplifier is clipped on the positive half-cycles by the first tube and on the negative half-cycles by the second. Positive grid bias for saturation limiting in the second stage is supplied from the B-plus circuit through $R1$. A large resistance, $R2$, which gives deliberately poor regulation of plate voltage, is used in series with the plate circuit of $V2$. When a signal is applied, the increase in negative bias caused by the negative-going pulses decreases the plate current. The result is that the voltage at the junction of the resistor and the tank circuit rises sharply.

(2) In the region of low plate voltage with no signal, the transconductance of the tube is low and the plate resistance also is reduced. These combine to reduce the noise output of the discriminator. The grid-cathode impedance of the second tube also is lowered by positive bias, which decreases the noise response of the tuned input circuit. When the signal makes the grid bias negative, the plate current drops, raising the plate voltage and reversing these conditions. The plate voltage change is approximately 50 volts with signal applied.

b. Cascade Limiter Disabling Audio Amplifier. This change in plate voltage can be used to disable the first audio stage. In the circuit shown in B, the control voltage across $R2$ is reduced by the voltage divider, R_a and R_b, and applied to the audio-amplifier grid. The cathode bias of the audio tube is adjusted so that the audio grid is sufficiently negative in respect to the cathode to remain cut off in the absence of signal. When a signal is applied, the voltage on the grid rises and the tube again can conduct. The bypass capacitor, C, is used to remove noise and hum from the audio grid.

c. Limiter Squelch Applied to Discriminator Diodes. The squelch voltage derived from the plate circuit of the second limiter can be applied directly to the diode plates of the discriminator, as in C. A positive voltage is placed on the cathodes by the voltage divider, R_aR_b. The diodes then are cut off because of more positive cathode voltage than plate voltage until a signal is received. With a signal present, the plates go positive in respect to the cathode, and the detector operates normally.

84. Discriminator-Derived Squelch

The circuit of figure 149 shows how squelch voltage can be derived from the discriminator circuit. The center of the load resistors, $R3$ and $R4$, of the discriminator develops a negative

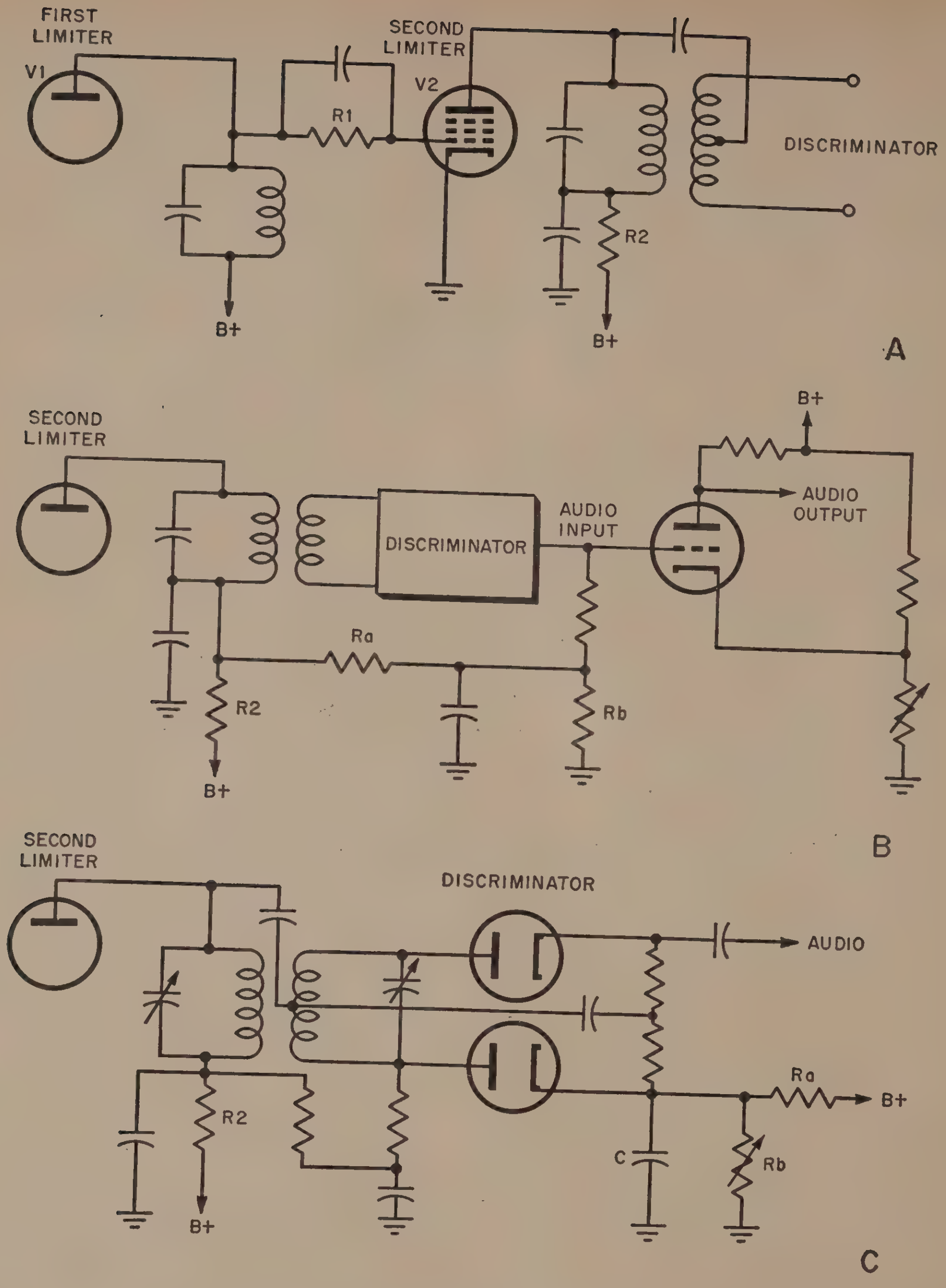


Figure 148. Limiter squelch circuits.

voltage when a signal is applied. This negative voltage applied through $R5$ and $R6$ is sufficient to cut off the squelch tube and permit a signal to pass through the audio amplifier. Without a signal, positive bias, applied to the junction of $R1$ and $R2$, is greater than the bias applied to the cathode of the squelch tube from $R11$ and $R12$. This causes the tube to draw current, dropping the plate voltage and cutting off the audio tube through $R7$.

85. Noise-Rectifier System

a. The circuit of figure 150 shows another type of squelch circuit. With no signal present, a certain amount of noise appears in the output of the discriminator. It is amplified by the noise

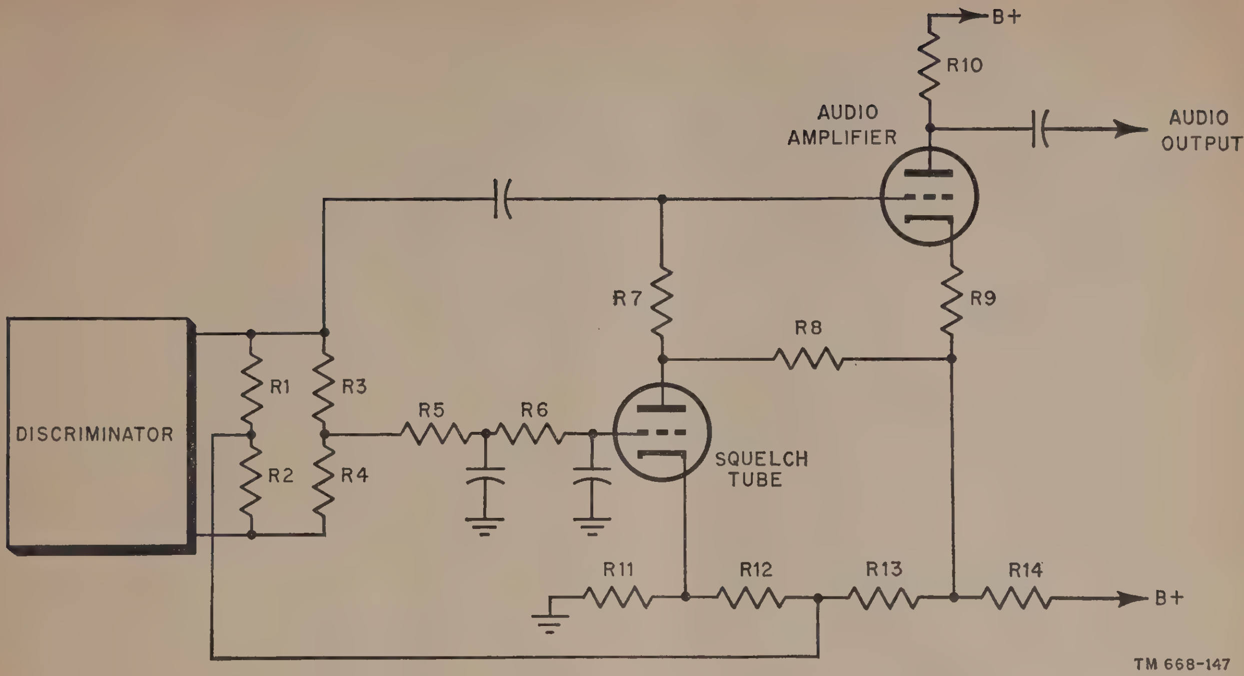


Figure 149. Discriminator-derived squelch.

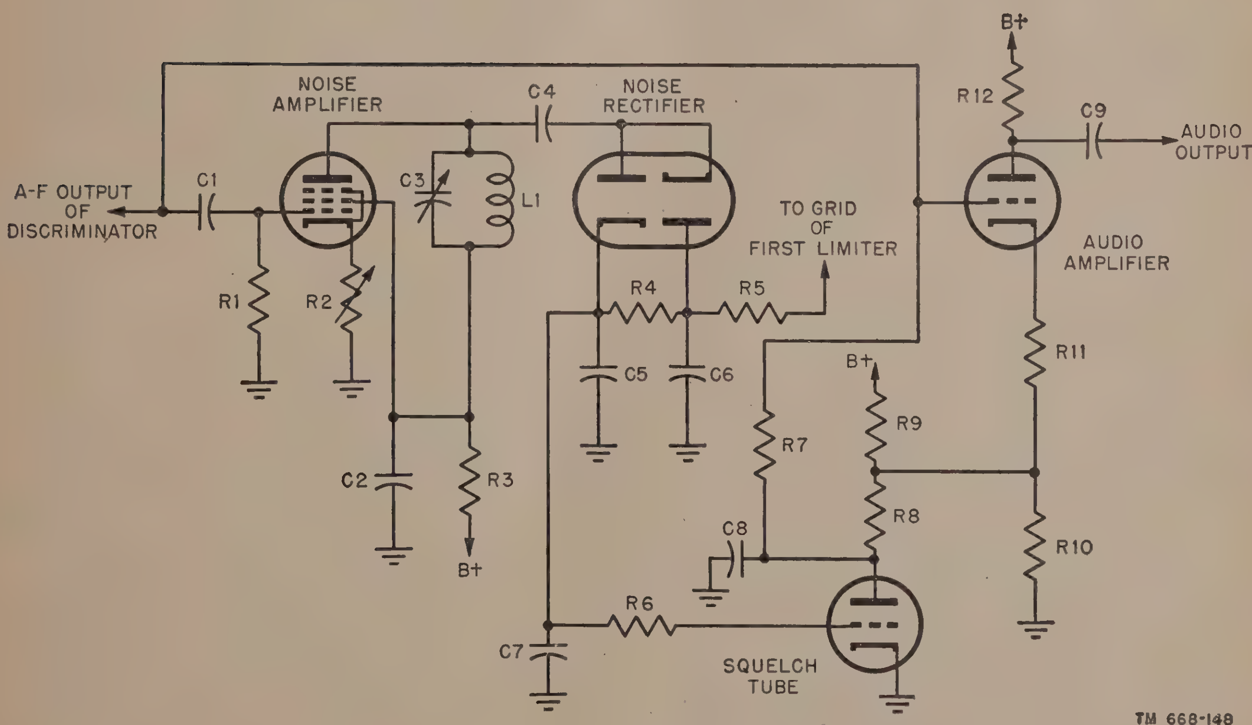


Figure 150. Noise-rectifier system of squelch.

amplifier, and the output is rectified by one section of the noise rectifier. The negative charge built up on $C6$ is applied to the grid of the first limiter through $R5$, reducing its gain. The positive peaks are rectified by the other half of the tube, and a positive charge is built up on $C5$ and applied to the grid of the squelch tube through $R6$. The squelch tube conducts heavily, cutting off the audio amplifier by making its grid highly negative.

b. A variation of this circuit applies the negative voltage generated by the first limiter grid directly to the squelch amplifier through isolation resistors. A diode is connected across one of these resistors to prevent rectified audio signals in the limiter grid circuit from reaching the squelch tube. In addition, the higher audio frequencies from the output of the discriminator are selected by a high-pass filter network and rectified in another diode and also applied to the squelch grid. With no signal present, a positive voltage is produced by the noise diodes, operating the squelch tube and cutting off the audio amplifier. In the presence of signal, the negative voltage from the first limiter grid cancels this voltage, permitting the audio amplifier to operate. Moreover, the noise-reducing property of f-m causes less positive voltage to be developed at the noise rectifier, increasing the speed of operation of the squelch circuit. This type of squelch can be adjusted to operate on very weak signals.

86. Squelch-Oscillator Circuits

a. The amplifier and voltage arrangement described above are unsatisfactory when the d-c supply voltage is likely to vary over a wide range. In portable equipment powered by batteries, a reliable squelch arrangement is needed which does not depend to so great an extent on voltage variations. The basis for the operation of such a circuit is the abrupt starting characteristic of a pentode oscillator. The circuit diagram of figure 151 shows a typical squelch circuit of this kind. A noise amplifier and rectifier are fed with noise derived from the discriminator output. The rectified noise voltage is applied to a d-c amplifier which is biased positively in the grid circuit to provide an adjustable squelch control. Only the high frequencies are applied to the noise rectifier because it is connected through a high-pass (R-C) circuit that presents a high impedance to any voltages in the speech range. With no signal received, the negative rectified voltage developed at the grid of the d-c amplifier is very large, and the tube is cut off. This produces a high screen voltage, which sends the squelch oscillator-rectifier tube into oscillation. The output is coupled to a second rectifier through a d-c blocking capacitor and is rectified to produce a high negative voltage. When this negative voltage is applied to the grid of the first audio tube, it drives the tube beyond cut-off, eliminating noise from the output.

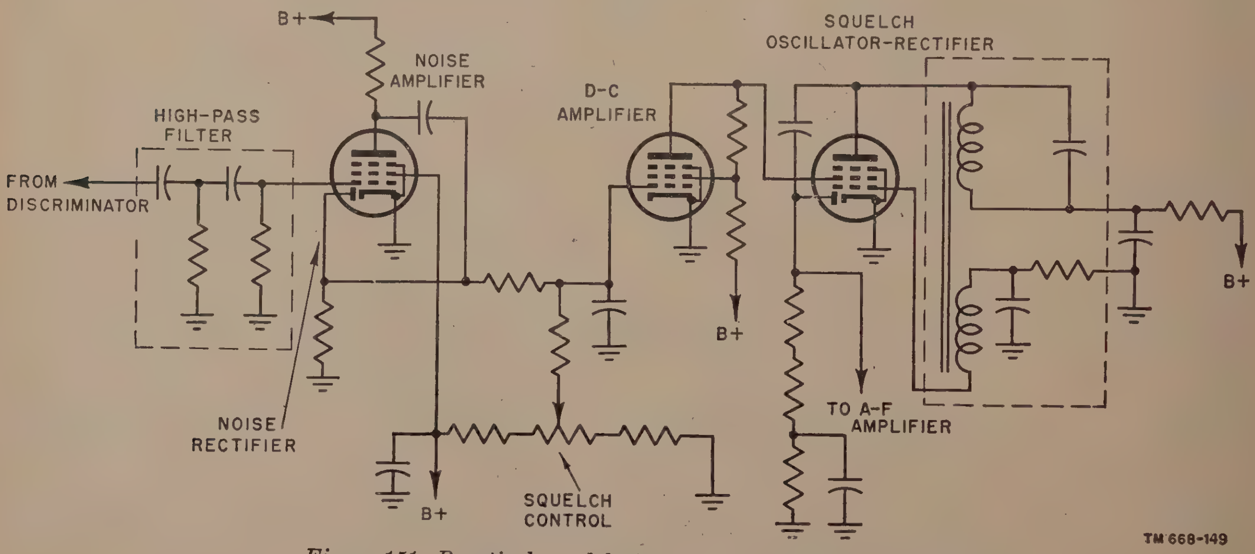


Figure 151. Practical squelch circuit for portable equipment.

b. With signal present, the noise-canceling properties of f-m remove voltage from the noise amplifier and rectifier. The d-c amplifier conducts, dropping the screen voltage, and the circuit no longer oscillates. The rectified voltage from the diode in the oscillator tube falls, and the audio amplifier can pass signals. The point at which the oscillator breaks into oscillation is relatively independent of the plate voltage. As the battery deteriorates, however, the squelch control must be readjusted. The circuit is better than the others that have been discussed, since falling supply voltage does not result in increasing the squelch level.

Section IX. AUDIO AMPLIFIERS, AFC CIRCUITS, AND TYPICAL RECEIVERS

87. Audio Amplifiers

a. The output of the f-m detector is usually about 1 volt. Audio power amplification therefore is necessary to operate loudspeakers, earphones, or other devices. The audio amplifier circuits used in most f-m receivers do not differ materially from their a-m counterparts, except in special equipment. The requirements of adequate response over the entire range of speech frequencies is easily met with simple transformer-coupled class-A amplifiers, using a single-ended beam tetrode or pentode. The inclusion of de-emphasis networks in the audio system is perhaps the only feature that distinguishes an f-m from an a-m audio system. They generally are incorporated between the output of the detector and the grid of the first audio amplifier where they operate at a low-voltage, high-impedance point, so that simple R-C combinations are adequate.

b. Occasionally, where communication is to be carried out under severe conditions, the frequency-response characteristic of the amplifier is shaped so that the lower speech frequencies (below 300 cps) are attenuated and the upper frequencies (above 3,000 cps) are accentuated. This contributes to the intelligibility of speech, since most of the consonant energy is in the upper range. It is possible to achieve a satisfactory result merely by restricting the values of the interstage coupling capacitors so that a rising frequency response is produced. Noise above 4,000 cps then can be attenuated by a small capacitor connected from plate to ground in one of the amplifier stages.

c. In some equipment, a wide a-f channel is required and the frequency response of the amplifier must be flat, from 50 to about 12,000 cps. These amplifiers are sometimes termed *high fidelity,* since a very wide frequency range is reproduced with low distortion. Resistance-coupled and transformer-coupled push-pull audio circuits with high-quality components are used. Inverse feedback, which consists of returning a portion of the output to the input in phase opposition, frequently is used. It results in the reduction of distortion and noise, and the extension of flat frequency response. The equipment which uses this high-fidelity arrangement is found most frequently in large fixed communication centers.

88. AFC Circuits

a. The use of automatic frequency control for receivers was mentioned in chapter 4 in connection with the necessity for keeping transmitters and receivers locked to the same channel. In general, advantage is taken of the discriminator in the receiver by utilizing some of its output to actuate a reactance tube associated with the high-frequency oscillator. This maintains the incoming signal in proper tune, regardless of drift in the receiver or transmitter oscillators. An afc circuit generally is used only when the detector is of the limiter-discriminator type, although it is possible to derive afc voltage from the tertiary coil of a ratio detector.

b. A d-c voltage is produced at the discriminator output if the signal is not exactly in tune with the center frequency. This voltage is positive or negative depending on whether the frequency is low or high. Such a condition takes place if the local oscillator in the front end drifts slightly. This, in turn, changes the relative value of the i-f, and appears as an off-center frequency in the fixed-tuned i-f amplifiers and the discriminator. The d-c voltage produced by the discriminator is applied to the grid of

a reactance modulator, changing the effective transconductance of the tube. The amount of reactance injected into the tuned circuit of the local oscillator then changes, varying the oscillator frequency in such a direction as to cancel out the drift which started the process. The response of the system must be sufficiently slow that variations in carrier frequency caused by modulation are not eliminated. Therefore, a long time constant is necessary in the network that feeds the reactance-modulator grid. In this way, only slow variations in the average frequency operate the reactance tube.

c. Where the transmitter and the receiver are combined in the same unit, the local oscillator of the receiver and the transmitter can be made interlocking and self-stabilizing by the use of various mixers and crystal oscillators. For example, in a typical receiver, the first i-f is approximately 5 mc and the operating frequency is 40 mc. Therefore, if the oscillator is operated on the low side of the received signal, a frequency of 35 mc is needed. The transmitter, by definition, operates on the same frequency as does the receiver. It must, therefore, produce output at 40 mc. A single local oscillator operating at 10 mc can be used for both units in the following way: The fourth harmonic of the oscillator is used directly in the transmitter, and the second harmonic (20 mc) is combined in a suitable mixer with the output from a crystal oscillator operating at 15 mc to give the needed 35-mc local-oscillator frequency.

d. To keep this complicated system interlocked with a distant transmitter requires only that the output of the receiver discriminator actuate a reactance modulator across the local oscillator to maintain the proper 5-mc difference. In doing so, the basic 10-mc frequency is maintained automatically when receiving. In the transmit position the receiver continues to operate, and some of the 40-mc output of the transmitter is received and operates the reactance system, keeping the transmitter locked to the receiver, which in turn already is locked to the distant station. There are other possible combinations of local oscillator, crystal oscillator, i-f, and transmitter frequencies that can be worked out with different arrangements of crystal oscillator and mixers in the same fashion.

89. Typical Receiver Circuits

a. Single-Conversion Superheterodyne.

(1) In the previous sections of this chapter, the operation of the individual circuits that go to make up a complete receiver have been discussed. When the entire receiver is at hand, it is customary to refer to the complete schematic diagram for needed information. At first glance, this diagram may appear complicated but part of the complex appearance can be attributed to the power supply and control circuit wiring, which often is drawn in a confusing manner and basically obscures the real functional circuits. To permit interpretation of a complete schematic, a typical single-conversion receiver is diagrammed in figure 152. The wiring for the filaments, plate supply, and control circuits has been eliminated for clarity. This receiver uses a single r-f stage of the grounded-cathode pentode type, *V*1, a pentode mixer, *V*2, a triode local oscillator, *V*3, three i-f amplifiers, *V*4, *V*5, and *V*6, a ratio detector, *V*7, and two stages of audio amplification, *V*8 and *V*9. Although a pentode, such as *V*1, does not result in the most efficient r-f stage, it is simple and stable in operation. The signal from the antenna is introduced by means of a tap on *L*. Capacitor *C*1 tunes the circuit to resonance, and is a part of the ganged tuning arrangement that simultaneously tunes the r-f, *C*1, the mixer, *C*3, and the oscillator, *C*2. The output voltage of the r-f stage is developed across *L*2, which is the primary of the transformer. The secondary is the tuned circuit, L3–C3, in the mixer grid circuit and is tuned to the same frequency as *C*1. Trimmer *C*27 adjusts the minimum capacitance in the input circuit and partially compensates for differing amounts of reactance that are injected across the input with different transmission lines and antennas.

(2) The input signal to the pentode, *V*2, mixer circuit appears across *L*3. Plate

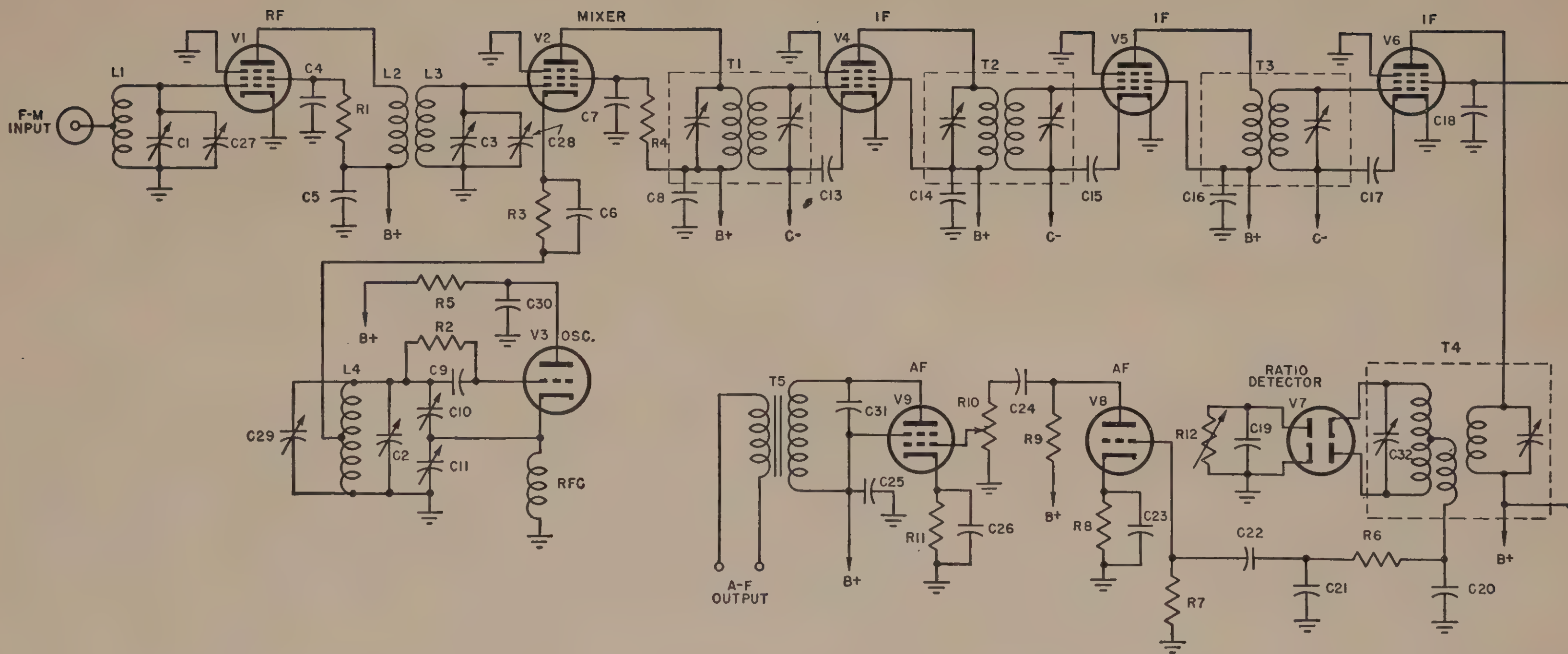


Figure 152. Typical single-conversion f-m receiver.

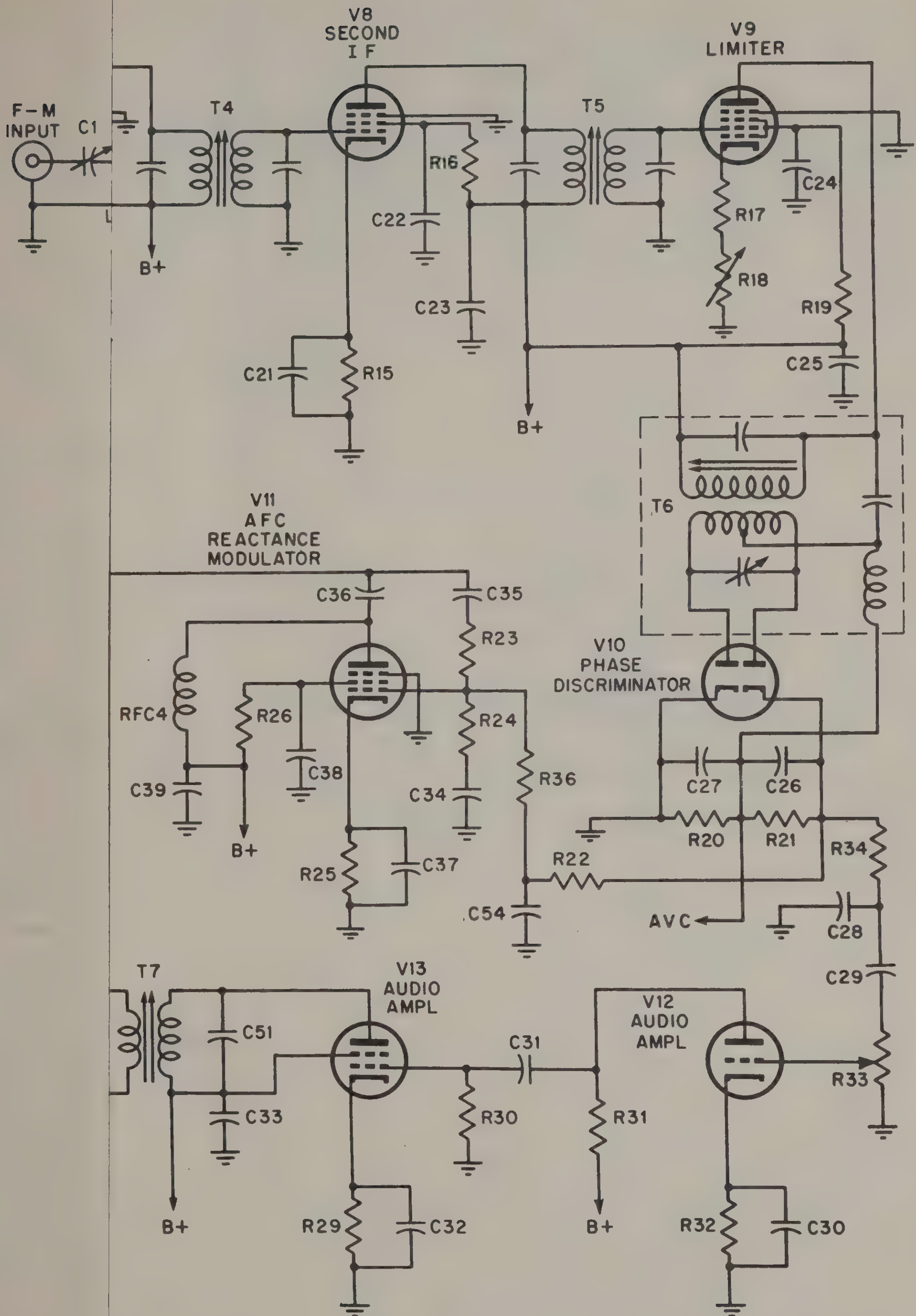
current is returned to the cathode through a tap on the lower end of the oscillator coil, $L4$. This injects some of the oscillator tank voltage into the cathode circuit through the r-f bypass, $C6$. The output tank circuit of the mixer is tuned to the intermediate frequency and forms the primary of $T1$, which is inclosed in a shielded can. The local oscillator is a Colpitts, with feedback provided by the cathode return to capacitors $C10$ and $C11$. Grid-leak bias is developed across $R2$ and $C9$. The actual tuning is done by the ganged variable capacitor, $C2$, which is across $L4$. The capacitor and resistor network, $C30$–$R5$, in the plate circuit of the oscillator, serves to ground the plate for r-f and prevent oscillator voltage from appearing in the power supply.

(3) The first two i-f amplifiers, $V4$ and $V5$, are sharp cut-off pentodes with separate cathode connections and use fixed bias so that the cathodes can be grounded directly. The grid and plate ground returns are made to the same point on the chassis through $C13$ and $C14$ in the first i-f and through $C15$ and $C16$ in the second i-f. The screen and the plate circuits have a common bypass capacitor, which provides some neutralization of the residual grid-plate capacitance. The primary of transformer $T3$ is untuned, but has sufficient inductance to be resonant at the operating frequency with the various stray capacitances. The capacitor across the secondary lowers the input impedance and reduces the detuning effects of a large signal applied to the grid of $V6$, the third i-f stage. The plate-load circuit of $V6$ is the primary of the ratio-detector transformer, $T4$. The time constant of the stablilizing circuit is adjusted for good a-m rejection by $R12$. Output voltage developed across the tertiary winding of the transformer and $C20$ is transferred to the first audio amplifier through the de-emphasis circuit formed by $R6$, $C21$, and blocking capacitor $C22$.

(4) The first audio stage, $V8$, is a conventional triode, resistance-coupled voltage amplifier with cathode bias furnished by $C23$ and $R8$. The output is developed across load resistor $R9$. The audio signal is coupled to the gain control, $R10$, through $C24$, which is made low in value to introduce some attenuation of the lower audio frequencies for improved intelligibility. The output voltage is developed across the transformer-coupled plate circuit of $V9$. High frequencies, above the useful speech range, are attenuated by $C31$. Actually, this capacitor and the inductance of the primary of $T5$ form a parallel resonant circuit at about 4,000 cps. Therefore, a greater load resistance is presented to the amplifier at this frequency and a greater output voltage is developed across it which serves to accentuate the consonants in the upper speech range. $C25$ serves as both a screen bypass and a plate return for the audio amplifier. Cathode bias is developed across $R11$ and $C26$ for class-A operation, with $C26$ made sufficiently large that the bypass action is poor at the lower audio frequencies. This introduces some out-of-phase voltage on the grid, effectively reducing the gain of the amplifier in this range. The result is a more effective cut-off at the low frequency than is obtained through a low value of $C24$ alone.

b. Typical Double-Conversion Superheterodyne.

(1) The complete schematic of a high-sensitivity, high-stability, double-conversion f-m superheterodyne is shown in figure 153. A direct-coupled, driven, grounded-grid, r-f stage, $V1$ is followed by a triode mixer, $V2$, which is cathode-coupled to a crystal overtone oscillator, $V3$. The first i-f signal is amplified by $V4$ and fed to the second mixer, $V5$, where it beats with the local oscillator signal from $V6$, producing the second i-f. The second i-f is amplified by $V7$ and $V8$ and applied



TM 668-151

296573 O - 54 Face p. 178

current is returned to the cathode through a tap on the lower end of the oscillator coil, $L4$. This injects some of the oscillator tank voltage into the cathode circuit through the r-f bypass, $C6$. The output tank circuit of the mixer is tuned to the intermediate frequency and forms the primary of $T1$, which is inclosed in a shielded can. The local oscillator is a Colpitts, with feedback provided by the cathode return to capacitors $C10$ and $C11$. Grid-leak bias is developed across $R2$ and $C9$. The actual tuning is done by the ganged variable capacitor, $C2$, which is across $L4$. The capacitor and resistor network, $C30$–$R5$, in the plate circuit of the oscillator, serves to ground the plate for r-f and prevent oscillator voltage from appearing in the power supply.

(3) The first two i-f amplifiers, $V4$ and $V5$, are sharp cut-off pentodes with separate cathode connections and use fixed bias so that the cathodes can be grounded directly. The grid and plate ground returns are made to the same point on the chassis through $C13$ and $C14$ in the first i-f and through $C15$ and $C16$ in the second i-f. The screen and the plate circuits have a common bypass capacitor, which provides some neutralization of the residual grid-plate capacitance. The primary of transformer $T3$ is untuned, but has sufficient inductance to be resonant at the operating frequency with the various stray capacitances. The capacitor across the secondary lowers the input impedance and reduces the detuning effects of a large signal applied to the grid of $V6$, the third i-f stage. The plate-load circuit of $V6$ is the primary of the ratio-detector transformer, $T4$. The time constant of the stablilizing circuit is adjusted for good a-m rejection by $R12$. Output voltage developed across the tertiary winding of the transformer and $C20$ is transferred to the first audio amplifier through the de-emphasis circuit formed by $R6$, $C21$, and blocking capacitor $C22$.

(4) The first audio stage, $V8$, is a conventional triode, resistance-coupled voltage amplifier with cathode bias furnished by $C23$ and $R8$. The output is developed across load resistor $R9$. The audio signal is coupled to the gain control, $R10$, through $C24$, which is made low in value to introduce some attenuation of the lower audio frequencies for improved intelligibility. The output voltage is developed across the transformer-coupled plate circuit of $V9$. High frequencies, above the useful speech range, are attenuated by $C31$. Actually, this capacitor and the inductance of the primary of $T5$ form a parallel resonant circuit at about 4,000 cps. Therefore, a greater load resistance is presented to the amplifier at this frequency and a greater output voltage is developed across it which serves to accentuate the consonants in the upper speech range. $C25$ serves as both a screen bypass and a plate return for the audio amplifier. Cathode bias is developed across $R11$ and $C26$ for class-A operation, with $C26$ made sufficiently large that the bypass action is poor at the lower audio frequencies. This introduces some out-of-phase voltage on the grid, effectively reducing the gain of the amplifier in this range. The result is a more effective cut-off at the low frequency than is obtained through a low value of $C24$ alone.

b. Typical Double-Conversion Superheterodyne.

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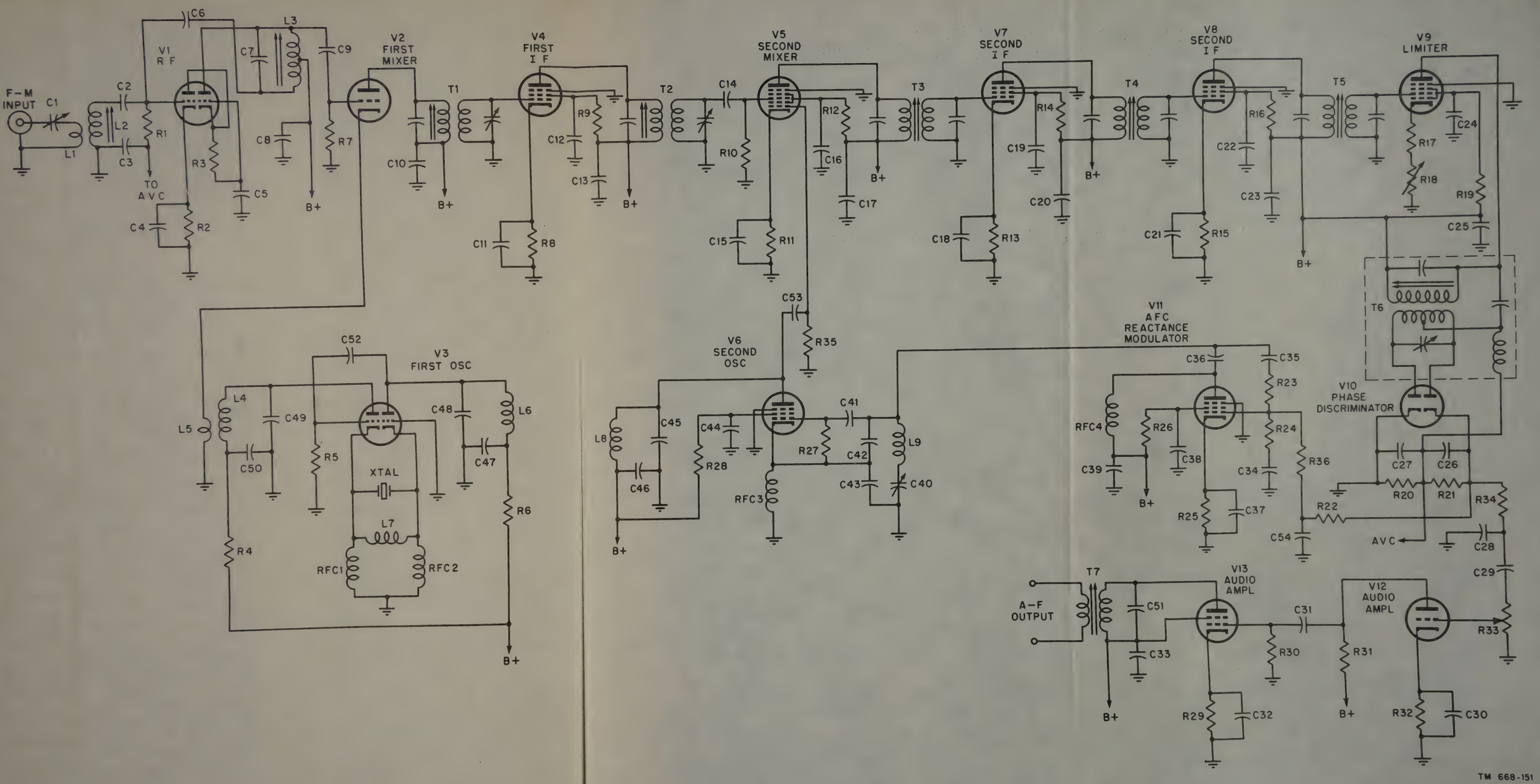
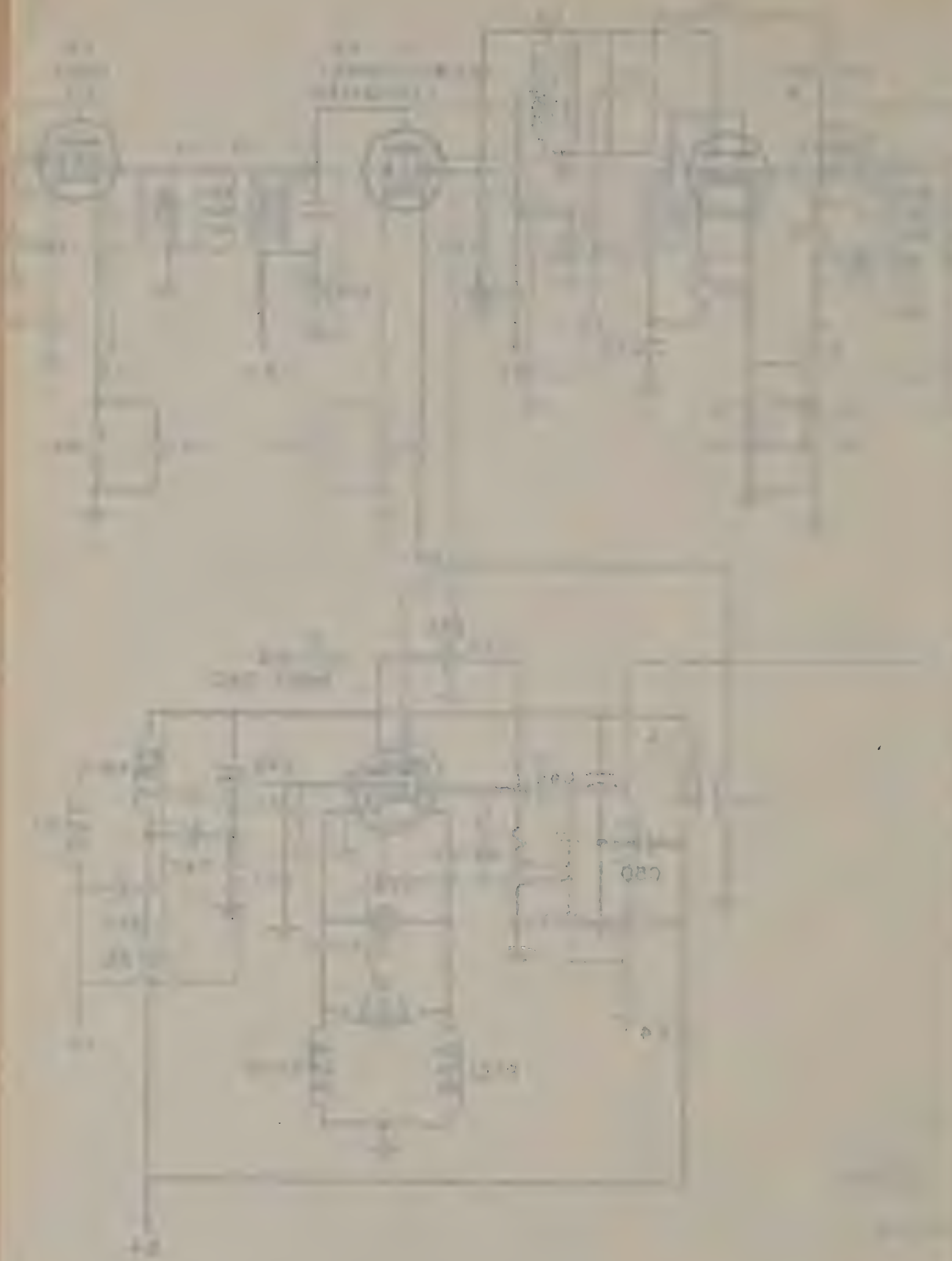


Figure 153. Typical double-conversion f-m receiver.



to the injection grid of a gated-beam tube, $V9$, operating as a limiter. The output of the limiter stage is transformer-coupled to the discriminator, $V10$, from which is derived the audio, afc, and avc voltages. The audio signal is fed to conventional audio amplifiers, $V12$ and $V13$, which are identical to those used in the single-conversion receiver. The afc voltage is applied to reactance tube $V11$, which changes the frequency of the second oscillator to produce a constant second i-f. The avc voltage is returned to the grid of the r-f amplifier.

(2) Since the first local oscillator is crystal-controlled, the input circuits of the r-f amplifier and mixer have a broad pass band. A double-tuned circuit formed by $C1$–$L1$ and $L2$ with stray capacitance is used at the grid of the first section of the driven, grounded-grid circuit. Since avc voltage is applied to the grid, it must be grounded for r-f by capacitor $C3$. The d-c voltage is fed through $R1$, which is isolated from the tuned circuit by $C2$. $C5$ and $R3$ develop the cathode bias for the grounded-grid section of the circuit. The direct-coupled circuit used permits a considerable reduction in the number of parts, and also in stray capacitance. The output signal is developed across the tuned circuit, $L3$–$C7$, which is center-tapped and grounded for i-f by $C8$. To improve the noise figure, the out-of-phase neutralizing voltage at the ungrounded lower end is coupled back into the input circuit by $C6$. Output voltage from the stage is capacitively coupled to the triode mixer, $V2$, through $C9$, and $R7$ serves as the grid-leak bias resistor. The crystal overtone first oscillator is inductively coupled to the cathode of the mixer by the tank circuit of $L5$. The output of the mixer is amplified in the second i-f amplifier, $V4$, and is coupled to grid 3 of the second mixer of the pentagrid converter, $V5$, through the i-f transformer, $T2$. The oscillations in the grid and screen circuits of the second oscillator, $V6$, are coupled to the plate through the electron stream, and appear across the output tank circuit, $L8$–$C45$. The output circuit is tuned to a harmonic of the actual oscillator frequency so that higher stability is obtained through isolation of the frequency-determining tank formed by $L9$, $C42$, $C43$, and $C40$. $R27$ and $C41$ serve as the grid-bias combination, and the r-f choke in the cathode permits the r-f voltage to develop across the feedback capacitors. The output of the second local oscillator is fed to grid 1 of the mixer through the $C53$–$R35$ network. The frequency of the oscillator is controlled by the reactance modulator of the afc circuit, $V11$, which injects inductive reactance into the tank circuit to compensate for any oscillator drift.

(3) Signal voltage from the second mixer is amplified by the conventional tuned i-f amplifiers, $V7$ and $V8$, with the needed selectivity provided by transformers $T3$, $T4$, and $T5$. The output of the secondary of $T5$ is applied to the first grid of a gated-beam-tube limiter, $V9$, which removes amplitude variations and impulse noise from the signal. The limited output is coupled through $T6$ to a conventional phase discriminator, which also produces the afc and avc voltages. Afc voltage is coupled to the grid of the reactance modulator through $R22$. The reactance modulator is prevented from responding to the variations in the signal by $R36$ and $C54$. $R34$ and $C28$ act as a de-emphasis circuit at the output of the discriminator. The remaining audio section of the receiver is the same as in the single-conversion receiver. Analyzing a complicated receiver circuit is a simple procedure when its component circuits are broken down into separate stages. With the inclusion of the control and power supply, the analysis is similar, although it sometimes is more difficult to see the interconnection of circuits and functions of the individual stages.

90. Single-Conversion F-M Receiver

A single-conversion receiver designed for use in vehicular or ground installations is shown in figure 154. Three separate receivers, alike except for the tuning range, cover the band of frequencies between 20 and 55 mc. The single-conversion receiver is small and compact, with all operating controls and cable connectors mounted on the front panel. Power is supplied by a storage battery and vibrator power supply. The stages include an r-f amplifier, a mixer, a local oscillator, four i-f amplifiers, a limiter, a discriminator, and a two-stage audio amplifier. The squelch circuit uses the pentode section of a pentode-diode tube as a tuned-plate tuned-grid oscillator. The rectifier section rectifies the output and applies a bias to the r-f amplifier tube to reduce its gain. It also applies bias to the audio-amplifier circuits, making them inoperative when no signal is applied. An audio output of 1 watt can be supplied for a speaker or 50 mw for phones.

91. Double-Conversion F-M Receiver

The receiver chassis for the 30- to 40-mc band shown in figure 155 is housed in the same cabinet as the indirect f-m transmitter shown in figure 101. This double-conversion receiver uses crystal oscillators for both low- and high-frequency conversion. A single r-f stage feeds the f-m signal to the mixer where it mixes with the signal produced by a crystal oscillator. The mixer output of the i-f is 4.3 mc and this output is amplified by an i-f amplifier and fed to a

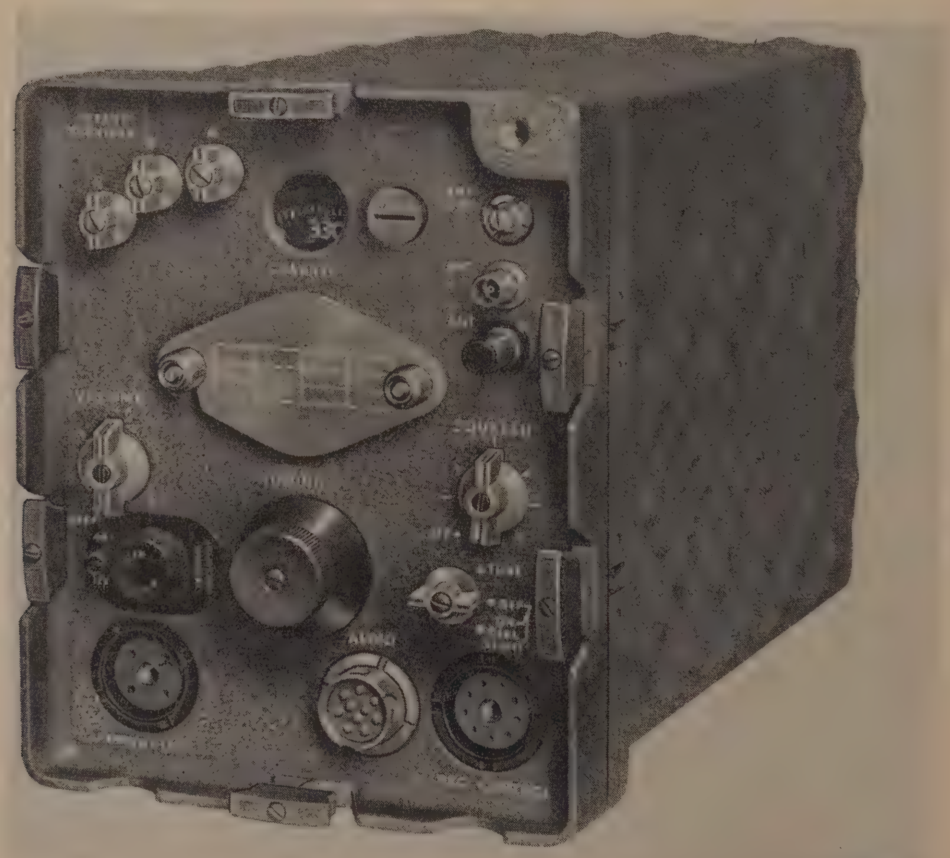


Figure 154. Typical single-conversion receiver.

triode-hexode mixer. The triode section is a crystal oscillator. The output of the hexode mixer section is at the low i-f of 455 kc. The low i-f amplifier feeds the limiter-discriminator section. A cascade limiter feeds the discriminator, which supplies an a-f signal to a two-stage audio amplifier. The output of the discriminator also is used to feed a noise amplifier and rectifier, which in turn inject a signal into the squelch circuit. The squelch tube cuts off the first a-f amplifier when no signal is present. If a signal is present, the first limiter renders the squelch circuit inoperative, allowing the first a-f amplifier to conduct normally. The final a-f amplifier provides sufficient output to drive a loudspeaker.

92. Walkie-Talkie

a. A versatile, battery-powered, f-m receiver-transmitter, designed to provide manpack communications for armored, artillery, and infantry units, is shown in figure 156. It also can be used in airplane and vehicle installations, or in semipermanent ground installations. The walkie-

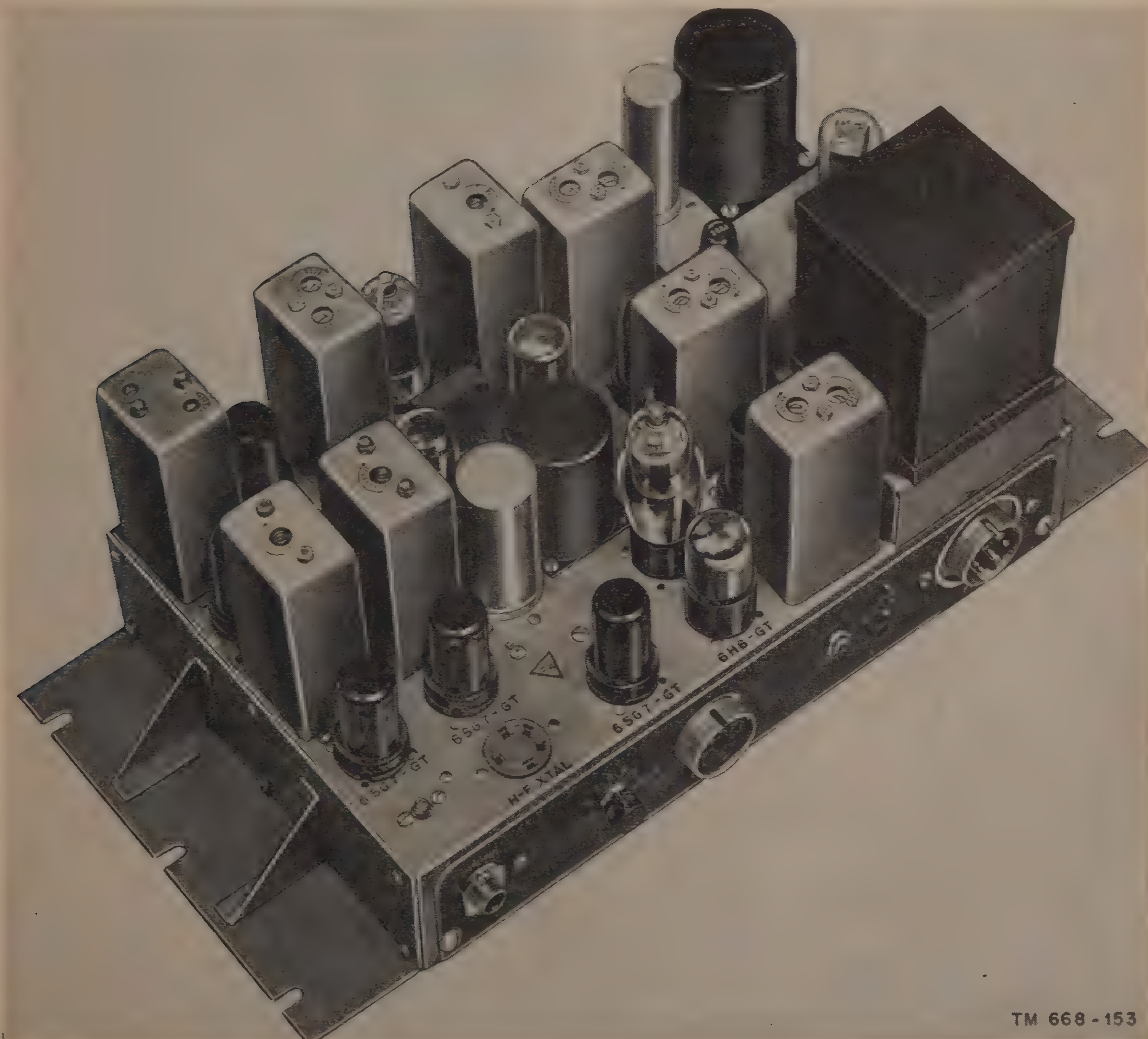


Figure 155. Typical double-conversion receiver.

talkie is available in three frequency ranges, 20 to 27.9, 27 to 38.9, and 38 to 54.9 megacycles, and the power output of the transmitter can be .9 watt, 1 watt, or 1.2 watts, depending on the model used. A single calibrated dial continuously tunes both the transmitter and the receiver. Direct f-m is used, and a built-in calibrator provides calibration points throughout the frequency range. The range varies from 3 to 12 miles. The transmitter consists of an electron-coupled oscillator and a nonlinear coil modulator plus an afc circuit. The afc circuit is controlled by a receiver interlock system which compares the frequency of the transmitter oscillator to that of the receiver local oscillator. Because the transmitter is not very powerful, the receiver accordingly is designed with high sensitivity. A .5-microvolt signal produces 2.5 milliwatts of output. The circuit consists of two r-f amplifiers, a mixer and a separate local oscillator producing an i-f of 4.3 mc, five i-f amplifiers, a phase discriminator using crystals as rectifiers, and a single stage of audio amplification. A squelch circuit and an i-f calibrator also are provided.

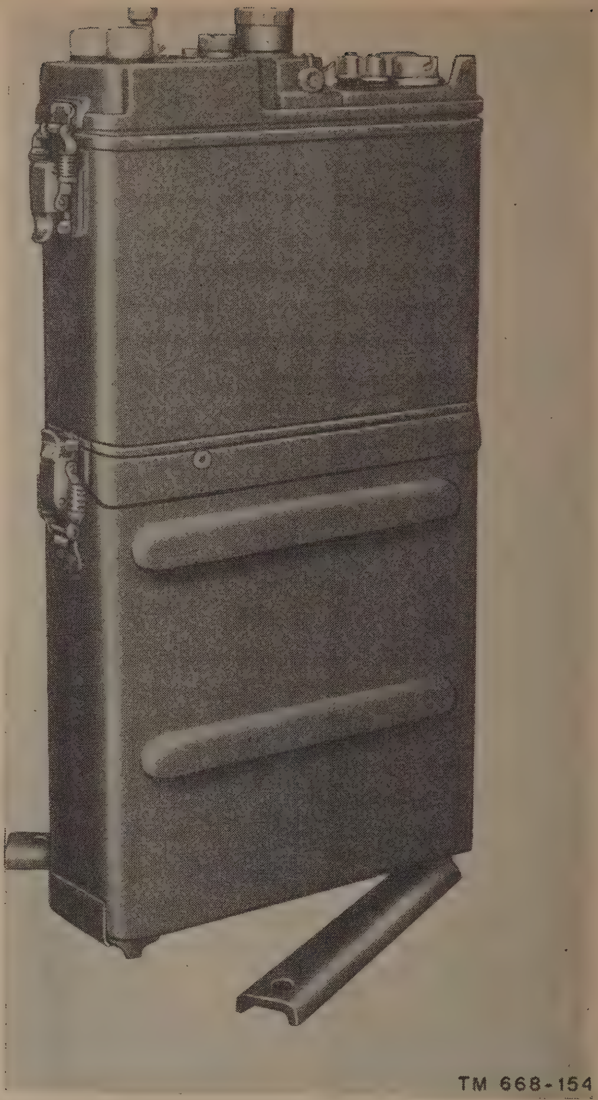


Figure 156. F-M walkie-talkie.

Section XI. ALINEMENT

93. General

a. All f-m receivers contain a large number of tuned circuits which must be adjusted correctly if the circuit is to function properly. Vibration, humidity, or temperature may cause the resonant circuits to drift off frequency in the field. Also, when a critical part is replaced, the associated resonant circuit may change frequency. The process of adjusting the tuned circuits of the receiver for optimum performance is called *alinement.*

b. The r-f, i-f, and detector circuits of any superheterodyne receiver must be alined to their proper frequencies. It is possible to accomplish a rough alinement by applying a signal at the antenna terminals and peaking the tuned circuits of the receiver until maximum output appears from the audio-frequency stages. This method is very unsatisfactory; however, in an emergency, with no servicing equipment available, the performance of the receiver can be restored to something approaching normal operation by this procedure.

c. The basic alinement procedure for any particular unit is given in detail in the manual which describes it. *Always refer to this manual for specific information.* This text indicates general principles applicable to most f-m receivers, and therefore must not be regarded as service information for any particular unit. The basic alinement procedure is to begin with the f-m detector circuit and tune the i-f amplifiers or limiters one by one, working toward the mixer, finally dealing with the r-f circuits.

94. General Alinement Methods

The two systems of alinement in common use are the *meter* method and the *visual alinement* method. The meter method uses a signal generator, which covers both the i-f and the entire r-f range tuned by the receiver, and a vacuum-tube, or high-resistance voltmeter (20,000 ohms per volt or better). The visual-alinement method uses an f-m signal generator that covers the r-f and i-f ranges, an oscilloscope, and sometimes a stable c-w signal generator. In general, an f-m receiver can be alined more quickly and easily and with far more accuracy by the visual method than by the meter method. The disadvantage of the visual method lies in the complexity of the test equipment and the difficulty of obtaining all of the necessary testing devices in the field. The meter method of alinement uses the variations of d-c voltage that take place in different parts of the receiver circuit when the tuning is changed. A steady carrier is applied from the signal generator to the circuit under test, and changes in voltage with tuning are observed on the meter. The visual method of alinement traces the actual response curves of the circuit under test on the screen of a cathode-ray oscilloscope. The tuned circuits are adjusted until the curves have the required amplitude and shape. The f-m signal generator sweeps through the frequency band covered by the stage under test, and the oscilloscope traces a curve that corresponds to the output of that stage in synchronism with the f-m generator.

95. Meter Alinement of Detectors

a. Discriminator.

(1) For meter alinement of the discriminator, a high-resistance voltmeter is needed as well as a signal generator capable of producing a stable i-f signal. When the secondary of the discriminator circuit shown in figure 157 is tuned to resonance at the desired center frequency, equal and opposite voltages are developed across $R1$ and $R2$. Consequently, there is no d-c voltage between points A and B. The tuning of the primary circuit of the discriminator affects the linearity of response. Therefore, the primary must be tuned for optimum linearity and the secondary must be tuned for correct center frequency.

(2) Connect the signal generator to the grid of the limiter tube which immediately precedes the discriminator. Connect the vacuum-tube or high-resistance voltmeter across either $R1$ or $R2$ ($M1$ in the figure). Adjust the tuning of the primary of the discriminator transformer for maximum deflection of the meter. The voltage developed across the resistor can be either positive or negative. When an ordinary meter reads backwards on the first reading, the leads must be reversed. When the meter reads maxi-

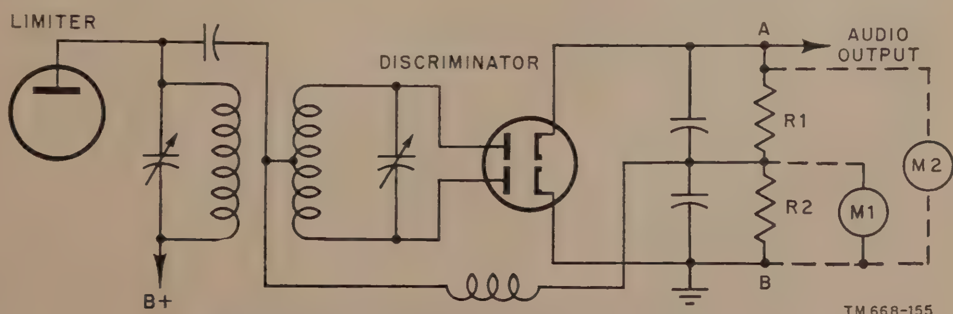


Figure 157. Discriminator alinement connections.

mum, the primary of the discriminator transformer is alined properly. When a meter with the mid-scale point at the zero-voltage mark is used, the polarity of connection is not important. Merely adjust for maximum deflection in either direction.

(3) Without changing the signal generator in any way, connect the meter across the entire load circuit (*M*2 in the figure). Since no voltage is developed across *A* and *B* when the tuning is correct, the meter reads zero when the secondary is adjusted properly. A zero-center meter is useful for this type of adjustment, since a positive and a negative voltage can be developed across the load resistors as the secondary is tuned, and the ordinary meter with zero at the extreme left of the scale is less convenient. Rock the tuning control of the secondary through zero several times, reducing the swing of voltage on either side each time, to insure accurate adjustment.

(4) The linearity of alinement can be checked easily with the meter. If the receiver must handle a deviation of 50 kc, shift the frequency of the signal generator 50 kc higher and then 50 kc lower than the carrier frequency. The meter across *A* and *B* should read the same total amount at either point, disregarding the change in polarity. With a zero-center meter, this adjustment can be observed easily as an equal deflection on either side of center for both frequencies. With an ordinary meter, the leads have to be reversed for one reading.

b. Ratio-Detector Meter Alinement.

(1) For meter alinement of the ratio detector, shown in figure 158, connect the signal generator to the grid of the last i-f tube and set it to the intermediate frequency desired. Connect the high-resistance voltmeter across the load resistor, *R*1, or the large load capacitor, *C* (*M*1 in the figure). Adjust the primary of the transformer for maximum deflection of the meter. This shows that maximum current is flowing through the load resistor and the capacitor.

(2) A symmetrical output circuit is necessary to tune the secondary. In some circuits, such symmetry already exists. When alining the circuit of figure 158, however, connect two 100,000-ohm resistors across the load capacitor, *C*. These resistors must be closely matched so that circuit symmetry is preserved. Connect the meter between the junction of the two resistors and the audio output end of the tertiary winding of the transformer. Tune the secondary for zero deflection of the meter. This shows that the secondary is exactly centered in the i-f pass band. Remove any resistors that have been

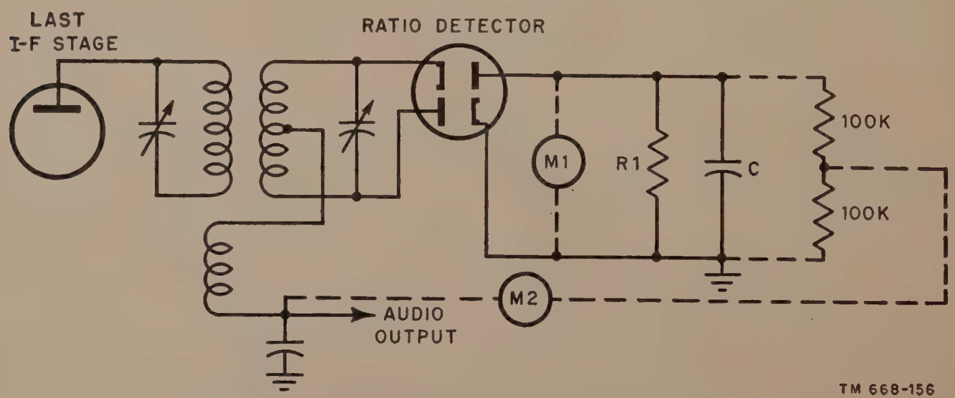


Figure 158. Ratio-detector alinement connections.

added. Linearity can be checked as in the discriminator by observing the swing of the meter on either side of center with a variation of the frequency of the signal generator.

(3) An alternative method for alinement of a ratio detector requires only an amplitude-modulated signal generator. Connect the signal generator to the grid of the last i-f stage and completely detune the secondary of the ratio-detector transformer by turning its slug or capacitor completely out. This unbalances the detector circuit and makes it sensitive to amplitude modulation as well as frequency modulation. Adjust the signal generator for about 30-percent modulation, and tune the primary of the detector transformer for *maximum* audio output. If an output meter is available, this adjustment can be made much more accurately by observing the meter deflection when it is connected across the audio output terminals. Be sure to advance the audio gain control so that a sufficient amount of signal can reach the meter, loudspeaker, or earphones.

(4) After the primary has been alined, adjust the secondary of the ratio-detector transformer for *minimum* audio output. This corresponds to maximum a-m rejection. This adjustment is critical and must be made slowly and carefully. Rock the tuning control back and forth to make sure that the minimum point is reached accurately.

c. Alinement of Locked-Oscillator Detectors.

(1) The alinement of a locked oscillator detector requires an a-m signal generator and an output meter. Connect the output meter across the audio output terminals. Turn the audio gain control to maximum. Ground the grid 1 of the detector tube; this stops oscillations and the tube functions as an a-m detector. Connect the signal generator to the grid of the last i-f tube and peak the detector transformer for maximum audio output. Then connect the generator to the second i-f tube grid and peak the next i-f transformer, continuing this process until the grid of the mixer is reached.

(2) Turn off the modulation in the signal generator and remove the ground from the grid of the detector tube. Short the quadrature circuit by connecting a jumper across the plate coil. The oscillator circuit continues to function. Adjust the oscillator trimmer capacitor until a heterodyne beat note is heard in the earphones or loudspeaker. This is the oscillator beating against the unmodulated carrier from the signal generator. Adjust the oscillator trimmer for zero beat, watching the output meter for an exact indication. The basic frequency of the oscillator now coincides with the frequency of the i-f signal generator.

(3) Remove the short from the quadrature circuit. Reduce the output from the signal generator to the point where the oscillator in the detector circuit is unable to lock in. Adjust the tuning of the quadrature coil so that the zero beat note is restored. This completes the alinement of the detector.

d. Alinement of Gated-Beam-Tube Detector.

(1) Since the gated-beam tube combines the functions of a limiter and a discriminator in one circuit, the alinement of the circuit is divided into two parts. Connect an *a-m* signal generator with a low percentage of modulation to the grid of the last i-f amplifier and an output meter to the audio output terminals. Turn the audio gain up fully. Short the quadrature circuit grid to ground, and decrease the output from the signal generator until a variation in the signal-generator output control produces a variation in the audio output. This setting insures an amount of i-f signal that is below the threshold of limiting. Tune the secondary and the primary of the detector input transformer for maximum audio output. As the amplitude of the signal increases with alinement, limiting begins to take place at the grid, and it is

difficult to determine sharp tuning points. When this difficulty occurs, reduce the output of the signal generator further, and repeat the peaking of the transformer.

(2) Increase the output of the signal generator to the point where limiting occurs as indicated by no further increase in audio output with increased signal-generator output. Increase the percentage of modulation and adjust the variable cathode resistor for minimum audio output. Repeat this threshold adjustment until further small changes in the cathode resistor setting do not affect the point at which the threshold of limiting begins. This completes the alinement of the limiter section.

(3) Connect an f-m signal generator with the required deviation to the grid of the last i-f amplifier. Remove the short from the quadrature circuit and tune the quadrature circuit for maximum audio output. This completes the alinement of the gated-beam tube.

96. Meter Alinement of I-F Stages

a. The meter alinement of the i-f stages in a receiver measures the voltage developed at the detector circuit as the i-f amplifiers are tuned. In the ratio detector, the meter is connected across the load capacitor, and in the gated-beam and locked-oscillator circuits, an audio output meter is used when the detector is altered so that it responds to a-m. When alining a receiver that uses a discriminator-limiter detector, connect the meter across the limiter grid resistor. If two resistors are used in series, connect the meter between the junction of the two and ground. As the signal to the limiter increases, the grid current increases, and consequently the voltage across the grid resistor rises.

b. Always work from the detector or the limiter toward the mixer. In the limiter-discriminator circuit, aline the discriminator first. If a dual limiter is used with tunable coupling between the two stages, aline the first limiter by connecting the meter across the grid resistor of the second limiter. Connect the signal generator to the grid of the first limiter. Then tune the interstage coupling circuit for maximum reading of the meter. Connect the meter to the grid resistor of the first limiter and proceed with the i-f alinement.

c. When all of the i-f amplifiers are known to be single-peaked—that is, when none of them are overcoupled—tune each secondary and primary, working toward the mixer from the first-limiter grid or from the grid of the last i-f stage, depending on the type of detector. Adjust each tuning control for maximum deflection of the meter. The signal generator must be unmodulated when the discriminator and ratio detector are alined, and amplitude-modulated for other detectors.

d. If one of the i-f amplifiers is overcoupled, the alinement procedure for that stage is slightly different from the procedure above. Using short leads, connect a low value of resistance (200 to 500 ohms) across the secondary of the overcoupled transformer. This serves to reduce the *Q* of the transformer, and the reduced *Q* suppresses the double-humped characteristic of the overcoupled circuit, with the result that there is a single broad resonant peak in the circuit. Observing the output meter carefully, tune to the exact center of this broad peak. The overcoupled circuit usually is the second i-f transformer in a limiter-discriminator type receiver. Disconnect the loading resistor when the stage has been alined.

97. Meter Alinement of R-F, Mixer, and Oscillator Stages

a. The alinement of r-f and mixer stages requires an accurately calibrated r-f signal source that can be tuned to frequencies in the low, high, and center portions of the receiver. The mixer and the oscillator must track over the desired range if the receiver is continuously tuned, and tracking adjustments may vary widely with the type of receiver. Consult the appropriate equipment manual for a particular piece of equipment.

b. The mixer trimmer and the oscillator trimmer are adjusted at the high frequency end of the tuning range for optimum gain and calibration. Set the receiver tuning control at the

designated alinement frequency. Connect a signal generator to the grid of the r-f amplifier. Set its frequency to the highest calibration point called for in the equipment manual. Vary the oscillator trimmer until the signal is heard in the output circuit, or until the meter in the detector circuit indicates maximum deflection. The generator is unmodulated for ratio and discriminator detectors, and amplitude-modulated for the others. Peak the mixer grid-circuit trimmer for maximum output. Tune the receiver to the designated low-frequency alinement point. Tune the signal generator to the low end of the frequency range and set it at the low-frequency calibration point. Adjust the oscillator padder until the signal is a maximum at the detector. Then return the signal generator and receiver to the high-frequency calibration point. Retune the oscillator trimmer, if necessary, to bring the dial calibration of the receiver into correspondence with the frequency of the generator. Repeat these steps until both the low and the high frequencies are on calibration without the need for touching either the oscillator trimmer or the padder. Set the signal generator to a point midway in the frequency range, and check the calibration. When these steps have been carried out carefully, the calibration should be correct.

c. Alinement of the r-f stage is done best with a noise generator, as described in the section on r-f amplifiers. Lacking a noise generator, a fair approximation is offered by the shorting technique described in that section. For the first rough initial alinement, antenna noise or a weak external signal supplied by the leakage from the signal generator or any other source will do. Peak the trimmer for maximum signal at the high end of the frequency range, and apply the noise-generator technique at the highest frequency to obtain the maximum signal-to-noise ratio.

98. Visual Alinement

a. Principles.

(1) In visual alinement, a cathode-ray oscilloscope and an f-m signal generator are used in place of the meter and the a-m signal generator. The actual response curves of the i-f and detector circuits are traced out on the scope by this method, whereas in the meter method of alinement only the maximum response points are known. Therefore, the visual method of alinement provides a more accurate means of adjustment and permits a very rapid grasp of exactly what is happening, since the actual behavior of the circuits with an f-m signal is visible.

(2) The oscilloscope is a device that responds to voltages; therefore, it always is connected in parallel with the circuit under test. The oscilloscope can be used to display currents, but first they must be passed through a resistor and then the voltage developed across the resistor is measured. The f-m signal generator is a standard signal generator with internal means for frequency modulation by either a sine wave or a triangular wave. The internal modulating signal is applied to the plates of the oscilloscope to provide a sweep voltage. For this reason, the device is called a *sweep* signal generator.

(3) Also useful in conjunction with visual alinement methods is an accurately calibrated signal generator (marker generator) which produces c-w signals to serve as directly visible frequency check points on the crt (cathode-ray tube) screen. Where a large number of receivers of a particular type are to be alined, crystal-controlled marker generators often are used for various critical frequencies, like the center of the i-f pass band and its outer limits.

b. Alinement of I-F and Limiter Stages.

(1) The input terminals of the scope provide for spot deflection both horizontally and vertically. For i-f alinement, the vertical-deflection terminals are connected across the limiter-grid resistor. This produces a voltage that is proportional to the amplitude of the signal input to that stage. The horizontal-deflection plates are connected

to the sweep voltages from the f-m signal generator. Connect the r-f output of the sweep generator to the grid of the last i-f stage. Adjust the frequency modulation for a deviation a few times greater than that of the rated deviation limits of the receiver. As the frequency swings through the i-f pass band, alining the tuned circuits changes the position and amplitude of the signal that appears at the limiter grid. Consequently, the variation of amplitude appears as changing vertical deflection on the screen of the cathode-ray tube. At the same time, the horizontal deflection changes in direct relationship with the sweep voltage. Since this also is the frequency modulation signal for the generator, the horizontal axis is effectively a frequency axis. Therefore, the picture displayed on the face of the tube when the signal is passed through an i-f amplifier is that of the familiar bell-shaped tuned-circuit response.

(2) As the tuning of the stage is varied, the response curve moves horizontally on the screen, in addition to changing in amplitude. Connect the marker generator to the grid of the last i-f stage. This causes a small wiggle to appear in the curve as the signal and marker generators beat against one another. If the marker generator is set to the intermediate frequency, the tuning of the circuits is adjusted to center the bell-shaped curve at the marker pip, as in A of figure 159. Similarly, two marker generators can be used to mark the limits of the i-f pass band, with a resultant trace on the screen that resembles that in B.

(3) Each of the i-f stages can be alined, connecting the sweep and marker generators to the grids of the preceding stages in turn. When a double-tuned overcoupled i-f circuit is used, the double-humped characteristic appears directly on the screen and can be centered at the marker frequency. There is no need to use any loading resistance. As additional stages are tuned, a broad, flat-topped characteristic can be developed that provides the least distortion and best off-channel attenuation.

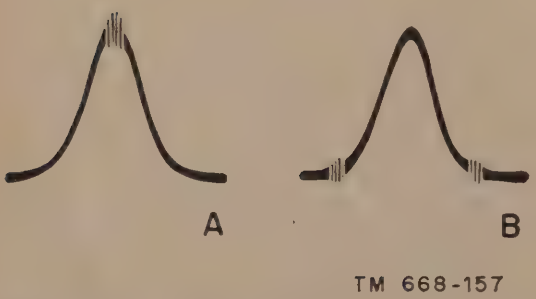


Figure 159. Use of marker frequencies in i-f alinement.

c. Alinement of Discriminators.

(1) Connect the vertical input of the oscilloscope to point A and ground in the discriminator of figure 157. Connect the f-m signal generator to the grid of the limiter immediately preceding the discriminator. Connect the horizontal-sweep voltage of the f-m generator to the horizontal input of the oscilloscope. As the signal on the horizontal plates varies, the spot on the screen traces out a line that corresponds to the amplitude of the modulating signal, and therefore to frequency. As the frequency changes, the instantaneous voltage developed across point *A* changes, and the spot is deflected either upward or downward, depending on its polarity and amplitude. When the discriminator is alined properly, the characteristic *S* curve is obtained, lying at an angle to the horizontal, and crossing the zero-voltage spot position.

(2) A marker generator connected to the limiter grid and tuned to the center frequency permits the development of the *S* curve in A of figure 160. When the marker generator is set at either the upper or the lower frequencies limit, the indications in B and C appear. As the secondary of the discriminator is tuned, the curve moves up or down, as in D. When the primary is

mistuned, the *S* curve is ragged, as in E. With correct tuning of the primary and secondary, the curve looks like that in F. First tune the primary for a smooth linear curve within the desired range, and then tune the secondary so that the line is centered on the marker frequency or on the center of the scope.

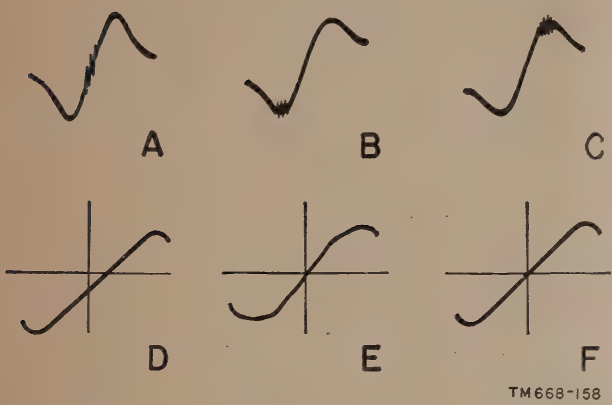


Figure 160. Use of marker frequencies in discriminator alinement.

d. Alinement of Ratio Detector.

(1) When visually alining receivers with ratio detectors, connect the vertical plates of the scope to the same point used in the meter method (fig. 158) or from the audio lead to ground. Connect the f-m signal generator to the grid of the last i-f tube.

(2) Detune the secondary trimmer. Peak the primary trimmer for a maximum single and symmetrical i-f curve. Aline the remainder of the i-f's, moving back a stage at a time toward the mixer. Then adjust the secondary of the ratio-detector transformer to give the S-shaped characteristic curve. A marker generator is helpful in setting the exact center.

(3) A second method of alinement starts similarly: Connect the scope and generator as above. Tune the primary for the single-peaked curve. Then tune the secondary for the *S* curve. Move the signal generator a step at a time toward the mixer, tuning each stage to obtain the largest symmetrical and linear *S* curve.

(4) Another method requires the opening of one lead of the load capacitor. Then the circuit operates as an a-m detector. The terminals of the scope must be connected across the load resistor. Adjust all of the i-f transformers for a symmetrical curve. Reconnect the capacitor, connect the scope to the meter position, or across the audio output to ground. The scope now shows the *S* curve. Adjust the secondary of the ratio-detector transformer for proper centering.

e. Alinement of Locked Oscillator.

(1) Connect the vertical input of the scope to the junction of the plate load resistor and the quadrature coil and ground. To aline the i-f stages, ground the oscillator grid. Connect the f-m signal generator to the grid of the last i-f tube and adjust the tuned circuits for the familiar single, peaked symmetrical response. Continue with the following i-f amplifiers, working toward the mixer.

(2) After the i-f amplifiers are alined, remove the short from the oscillator grid, tune the generator to the i-f center, turn off the f-m, and short the quadrature network. The signal generator and scope remain where they were. Adjust the oscillator trimmer for zero beat, as indicated by minimum signal on the oscilloscope. Remove the short from the quadrature circuit, turn on the f-m in the generator, and set the deviation for the rating of the receiver. Adjust the quadrature coil for linear response on the oscilloscope screen (fig. 161). In A, the quadrature circuit is misalined. In B, the alinement is correct. In C, both the quadrature circuit and the last i-f transformer secondary circuits are misalined. Marker generators are not needed for the center frequency of the

locked oscillator. Frequently, it is easier to use meter methods of alinement for the r-f stages and visual alinement for the i-f and detector circuits. This requires the use of a considerable amount of equipment, but saves time when a large number of receivers of the same type must be alined.

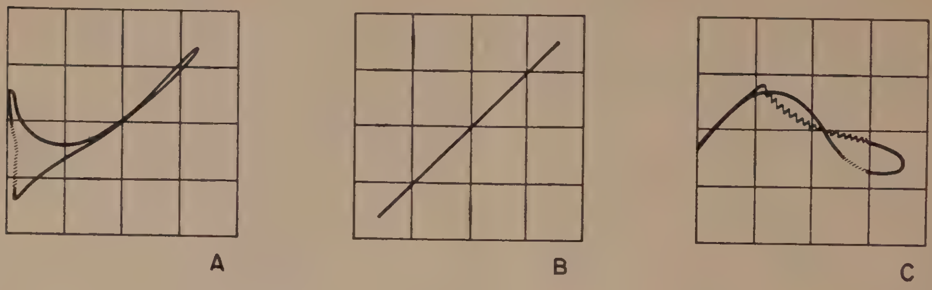


Figure 161. Scope patterns for locked-oscillator detector.

Section XII. SUMMARY AND REVIEW QUESTIONS

99. Summary

a. F-m receivers generally are of the superheterodyne type.

b. The f-m signal voltage in a superheterodyne is amplified and combined with the voltage from a local oscillator in a mixer. The difference-frequency output is fed to an i-f amplifier.

c. The output of the i-f amplifier is converted into audio voltage in an f-m detector. The audio signal is amplified and applied to speakers, headphones, or other audio devices. In the absence of signal, a squelch circuit can be used to prevent circuit noise from reaching the output.

d. For greater selectivity, double-conversion superheterodynes are used. The output of the first i-f amplifier is converted to lower second i-f.

e. The r-f amplifier receives the lowest level signal in the receiver from the antenna and amplifies it before it reaches the mixer. The critical features in the performance of the r-f amplifier are image rejection, noise figure, and attenuation of radiation from the local oscillator.

f. The noise figure of the r-f amplifier depends on its bandwidth, the type of circuit, and the tube used. The noise performance of the receiver depends almost entirely on the noise figure of the r-f amplifier, if its gain is high. The image rejection depends on the selectivity of all tuned circuits encountered by the signal before reaching the grid of the mixer.

g. The mixer stage operates as a low-level modulator, with the local oscillator acting as the carrier and the signal as the modulation voltage.

h. The voltage gain of a mixer depends on its conversion transconductance, which is approximately the ratio of the i-f plate current to the r-f grid voltage.

i. The local oscillator signal can be injected on the same or different tube electrode compared to that to which the signal is applied. The type of oscillator injection determines the extent of its radiation, interaction between oscillator and mixer circuits, and loading of the mixer input circuit.

j. Multigrid mixers are used frequently for the second i-f sections of double-conversion receivers.

k. Converters are oscillators and mixers combined in the same tube envelope. The electron stream of the tube serves for both sections with a large number of grids performing the different functions necessary.

l. At very-high and ultrahigh frequencies, crystal mixers can be used. These units have conversion loss instead of gain, and are seldom

preceded by r-f amplifiers. Their noise performance depends on the following i-f stage. They are used with a very high i-f to obtain good image rejection and reduction of local-oscillator radiation.

m. Lumped-constant oscillators occasionally are used in the first oscillator stage of double-conversion receivers, and almost always in the oscillator stage of single-conversion receivers. Crystal oscillators characterized by high stability are used in the first oscillator stage of many double-conversion receivers and in the second oscillator stage of others. Distributed-constant oscillators are used in the oscillator stage of very-high-frequency receivers where the tuning range needed precludes the use of crystal oscillators.

n. Crystal oscillators used in f-m receivers are of two types—harmonic multiplier and overtone. Overtone oscillators operate on only one frequency, and harmonic oscillators produce their principal output on one frequency and lesser amounts of power at other harmonics.

o. The *front end* of the receiver consists of the r-f amplifier, oscillator and mixer stages. The design of the front end takes into consideration the need for automatic tuning. Various mechanical systems, electrical switching, or turret circuits can be used.

p. The i-f amplifier provides all of the selectivity and most of the sensitivity of the f-m receiver. The selectivity depends on the tuned circuits used in the i-f voltage amplifiers.

q. Transformer-coupled circuits are used in i-f amplifiers for high-gain, broad-frequency response, and sharp adjacent-channel selectivity. The stability of the i-f amplifier depends on the shielding, parts placement, and amount of feedback through the grid-plate capacitance.

r. Limiter circuits generally are i-f amplifiers operated with low screen voltage so that they overload easily on strong input signals. They eliminate a-m, and also some of the noise that appears along with the desired signal.

s. Discriminators are characterized by fair linearity, ease of alinement, and easily derivable afc and avc voltages.

t. Ratio detectors require no limiters, since they are relatively insensitive to a-m. The sensitivity is higher than the limiter-discriminator combination, but the circuit is more difficult to aline and has less rejection of impulse noise.

u. The synchronized single-stage locked-oscillator detector rejects interference from signals on the same channel as the desired signal if these signals are only slightly weaker. It is sensitive, but difficult to aline, and has poor immunity to some types of noise.

v. The cycle-counting detector requires no alinement and produces extremely linear output. It is used largely in frequency-meters, where the large number of stages of amplification, limiting, and conversion required for high sensitivity are not needed.

w. The gated-beam tube operates as an excellent limiter, and as a fair limiter-discriminator combination. The response is linear over a wide range, with good rejection of a-m. It is one of the easiest of all detectors to aline and has sensitivity comparable to the standard limiter-discriminator.

x. Squelch circuits used in f-m receivers operate either from the change in voltage in limiter plate circuits, or from change in noise level at the detector output. These voltage changes trigger the squelch circuit, which is normally biased to cut off the first audio stage. With applied signal, the squelch tube permits the first audio stage to conduct.

y. The afc system used in some receivers takes a voltage from the discriminator detector and applies it to a reactance modulator. This modulator then changes the frequency of the local oscillator to maintain proper tuning.

100. Review Questions

a. Explain the difference between a single- and a double-conversion superheterodyne, describing the functions of the various parts of each circuit.

b. What stage governs the receiver sensitivity more than any other?

c. Where is selectivity obtained?

d. Why does the r-f amplifier play an important role in the rejection of images?

e. What are the three basic functions of the r-f amplifier?

f. What is the effect of the input bandwidth of the r-f amplifier on the noise performance of the stage?

g. Describe the relative performance of different tube types used for r-f amplifiers.

h. What is the effect of a high-*Q* input circuit in an r-f amplifier?

i. Describe the function of the input circuit in respect to impedance matching. What is the difference between an unbalanced and a balanced input circuit?

j. What is the disadvantage of a grounded-cathode triode amplifier? How does its gain compare with the grounded-grid circuit?

k. What is the advantage of a push-pull triode r-f amplifier over a single-ended pentode?

l. What is the main advantage of the cathode-coupled amplifier?

m. Describe two methods of neutralizing the driven grounded-grid amplifier.

n. What is the basic principal underlying frequency conversion?

o. What is the difference between a mixer and a converter?

p. How are mixers classified in regard to local-oscillator injection?

q. How are spurious responses generated in the mixer? How does the choice of i-f affect the production of spurious responses?

r. Why does a mixer produce more noise than the same tube used as an amplifier?

s. What is the main disadvantage of the triode mixer in regard to the stability of the stage?

t. What is the principal advantage of the push-push triode mixer?

u. In what way is the pentode mixer superior to the triode mixer?

v. What are the characteristics of the pentagrid mixer tube that make it very desirable for the mixer stage in the low-frequency section of a double-conversion receiver?

w. What is the principal reason for the use of the crystal mixer? What is the effect of the oscillator injection on the performance of the crystal mixer?

x. Why does the noise performance of the crystal mixer depend on the noise figure of the following i-f amplifier?

y. What is the principal requirement for the high-frequency oscillator in an f-m receiver?

z. What are the principal causes of frequency drift in oscillators and how are they overcome?

aa. Why are crystal-controlled circuits most useful as high-frequency oscillators?

ab. On what frequencies other than the fundamental does the crystal operate in a harmonic crystal oscillator?

ac. What principal disadvantage of the harmonic oscillator is overcome in the overtone circuit?

ad. Why are overtone oscillator circuits used in preference to L-C oscillators at very-high frequencies?

ae. What is the purpose of automatic tuning in f-m receivers?

af. When are transmission-line elements made a part of the tuning system of an f-m receiver?

ag. What are the principal means of tuning an i-f transformer?

ah. Describe two procedures used to obtain high adjacent-channel selectivity?

ai. Describe the major causes for instability in i-f amplifiers.

aj. Where extremely high gain is required of an i-f system, what means can be used to isolate the input from the output circuits?

ak. Describe the generation of the grid-bias voltage in a pentode grid-bias limiter.

al. Why is the output of a limiter substantially independent of the input signal over a very wide range?

am. Why are cascade limiters often used?

an. Why is it undesirable to have any appreciable gain in the limiter stages?

ao. What is the principal disadvantage of the double-tuned discriminator?

ap. How is this overcome in the phase-discriminator circuit?

aq. What is the principal disadvantage of a discriminator in respect to adjacent channel selectivity?

ar. How does the ratio detector provide immunity from a-m superimposed on the f-m signal?

as. How does the sensitivity of the ratio detector compare with the limiter-discriminator?

at. How do the side responses of the ratio detector differ from those of the discriminator, even though the tuned-circuit construction is very similar?

au. Tell why the quadrature circuit is necessary in the single-stage locked oscillator detector.

av. In what respects does the single-stage locked oscillator resemble a reactance modulator?

aw. What is the main reason for using a cycle-counting detector in frequency meters for f-m applications?

ax. Describe the passage of the electron stream through the gated-beam tube.

ay. Why does the gated-beam tube have such a very sharp cut-off characteristic?

az. What are the principal advantages of the gated-beam tube as a limiter?

ba. Describe the gating action of the two grids of the gated-beam tube, and show how audio voltage is derived.

bb. Give the purpose and function of a squelch circuit.

bc. Describe three methods for deriving a voltage to operate a squelch tube.

bd. Why is the squelch-oscillator circuit used in battery-operated equipment?

be. How does the afc circuit in some receivers resemble the circuits used in transmitters?

INDEX

—U. S. GOVERNMENT PRINTING OFFICE: 1957

7
15
10
32